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# HIGH SPEED PHOTOGRAPHY BY MEANS OF THE IMAGE CONVERTER

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*When Holst and his co-workers invented the image converter, the fascinating thing about this device was the possibility of evading Stokes' law in fluorescence and, therefore, transforming infrared images into visible ones. In recent years, this application of the image converter has grown rather commonplace and new possibilities have been discovered, one of the latest and most spectacular being the application of the image converter as an extremely fast acting photographic shutter, allowing exposures of less than $10^{-7}$ second.*

## Introduction

The basic idea of the image converter as illustrated in *fig. 1*, is the transformation of an optical image into an image formed by electron rays, this then being reconverted into an optical image [1]). The first transformation is effected by means of a semi-transparent photocathode, the second by means of a fluorescent screen. The electrons emitted from the photocathode are focused to an image on the screen by a suitable electron-optical system, in which the screen is part of the anode.

The device has found large-scale application for rendering infra-red images visible to the eye [2]). In this case use is made of the fact that light emitted by the fluorescent screen can have shorter wavelengths than the radiation forming the original image. Another important application of the device is the X-ray image intensifier [3]). This is based on the possibility of imparting energy to the electrons from an accelerating electric field, thus intensifying the image, whilst the brightness on the screen can at the same time be increased by using a *de*magnifying electron-optical system.

The application of the image converter as a fast acting photographic shutter, that will be dealt with in this article, is due to still another of its inherent properties: the electron rays constituting the intermediate image can easily be controlled, i.e. suppressed and released or deflected, in which controlling action practically no inertia is involved. With the aid of an image converter tube specially developed for the purpose of high speed photography in the Mullard Research Laboratories, the ME1201 (*fig. 2*), single shots with exposure times of about $3 \times 10^{-8}$ sec have been taken.

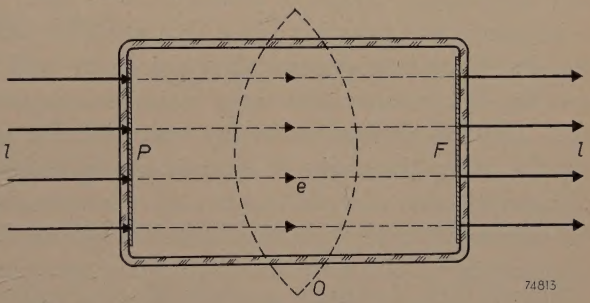

Fig. 1. Basic design of an image converter. *P* photocathode, *F* fluorescent screen, in evacuated vessel; *l* light rays, *e* electron rays; *O* symbolizes an electronoptical system.

For a comparison it should be mentioned that electronic flash tubes deliver flashes of light of duration between appr. $2 \times 10^{-4}$ and $2 \times 10^{-6}$ sec (2 μsec). With a flash duration of 2 μsec, obtained with the LSD2, the fastest of the Mullard range of

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[1]) G. Holst, J. H. de Boer, M. C. Teves and C. F. Veenemans, An apparatus for the transformation of light of long wavelength into light of short wavelength, Physica (The Hague) **1**, 297-305, 1934.

[2]) See for instance: T. H. Pratt, Electronic Eng. **20**, 274, 314, 1948.

[3]) See for instance: M. C. Teves and T. Tol, Electronic intensification of fluoroscopic images, Philips tech. Rev. **14**, 33-43, Aug. 1952.

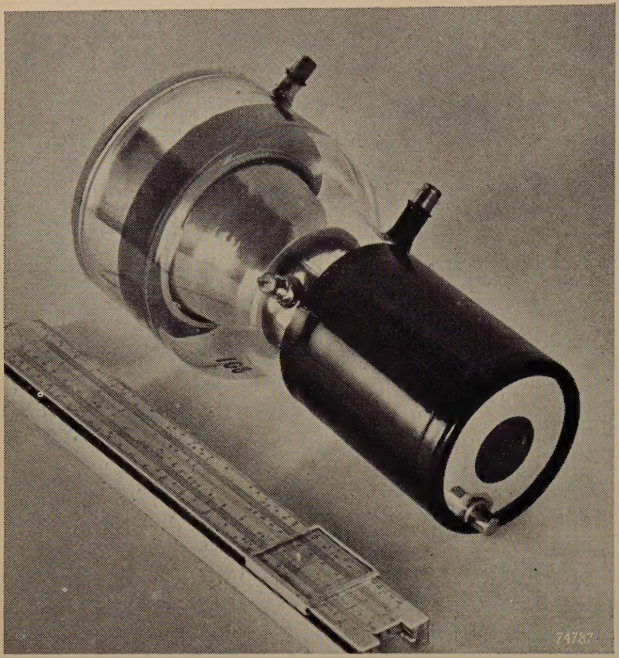

Fig. 2. The Mullard ME1201 image converter tube for pulsed operation.

photoflash tubes 4), the most extravagant requirements of amateur and professional photography in stopping fast action are by far exceeded. In the field of research, however, many instances of very rapid events are encountered which require an exposure time of this order or even less in order to obtain sharp pictures. Some examples are: the behaviour of projectiles impinging on obstacles, the vibrations and mechanical failure of rotors at high rates of revolution, fuel injection from high pressure jets, the distribution of stresses in dynamically loaded models, etc. In several of these cases, an exposure of 2 μsec will already be far too long. In other instances, as in the study of combustion, spark and discharge phenomena, detonation of explosives, etc. the objects of the investigation are self-luminous. In these instances an illuminating flash in any case cannot be the answer to the photographic problem, but an extremely rapid shutter is needed to permit the use of exposure times that are short as compared with the light emission period of the event under examination. The exposure must then be controlled by means of a synchronization method similar in principle to the well known synchronization used in conjunction with normal flash bulbs and normal photographic shutters, but different in that extremely short delay times and triggering times are required. Special synchronizing circuits making use of hydrogen thyratrons have been developed to this end.

A brief account of the development of the ME1201 image converter mentioned above and of the synchronizing circuits will be given in this article 5).

## Set-up

The set-up for applying the image converter as a photographic shutter is shown schematically in *fig. 3*. A sharp and bright image of the event or object to be photographed is produced on the photocathode of the image converter tube by a large aperture photographic lens of a type used in modern miniature cameras. At the other side of the tube a camera suited for reproduction work is permanently focused on the fluorescent screen. Fig. 3 also shows the triggering and synchronizing circuits that are connected to the image converter tube in order to let the electronic image pass only during the prescribed exposure time and to synchronize the exposure with the event under examination. In cases where the object is not self-luminous but is lighted for instance by an electronic flash tube, the circuits must also trigger this light source at the right moment.

The main part of this article will be concerned with the taking of single shots only. The problems involved in taking a number of pictures in rapid succession will be briefly dealt with in the last section of this article.

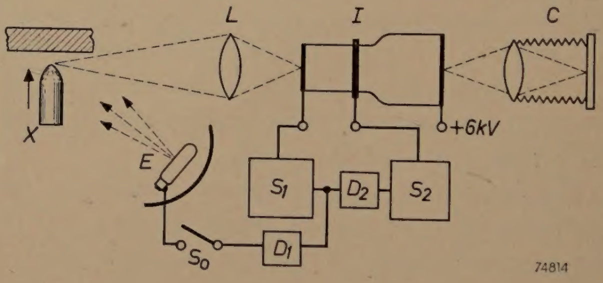

Fig. 3. Set-up for using the image converter as a high speed photographic shutter. *X* event to be photographed, *E* illuminating light source, *L* lens, *I* image converter tube, *C* photographic camera permanently focused onto the screen of *I*. The light source and the image converter are triggered and controlled in the correct intervals by means of switching and delay circuits $S_0$, $S_1$, $S_2$ and $D_1$, $D_2$.

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4) G. Knott, High-intensity flash-tubes, Photographic J. **89B**, 46-50, 1949.

5) A first and rather extensive account was given by the same authors in J. Brit. Inst. R.E. **11**, 505-517, 1951 (Nov.). Cf. also: Electronic Eng. **24**, 302-307, 1952 (July) and Photographic J. **92B**, 149-157, 1952 (Sept.-Oct.). The circuits to be used in conjunction with the image converter were extensively dealt with by M. S. Richards, An industrial instrument for the observation of very-high-speed phenomena, Proc. Inst. El. Eng. **99** part III A, 729-746, 1952 (No. 20). — It may be mentioned that the idea of using the image converter as a high speed photographic shutter seems to have been put forth first by R. T. Bayne in a patent application filed as early as 1943 (U.S. Patent No. 2, 421, 182, May 27, 1947). Several workers independently developed the same idea, one of the foremost being J. S. Courtney-Pratt (Research **2**, 287, 1949).

*Fig. 4* may serve to illustrate the quality of the picture obtained in a single shot of exposure time $2 \times 10^{-7}$ sec.

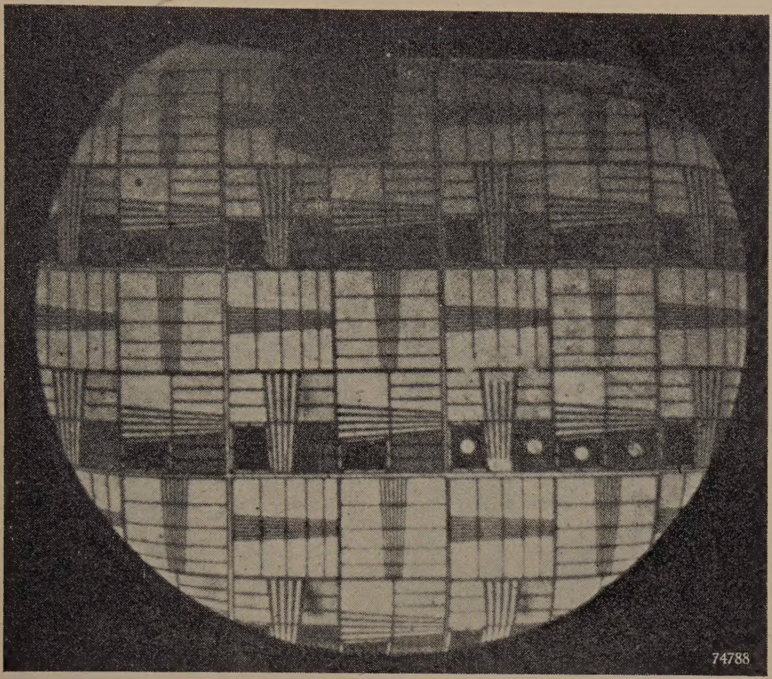

Fig. 4. Picture of a test chart recorded from the fluorescent screen of the image converter tube ME1201 with an exposure time of $2 \times 10^{-7}$ sec.

## Design of the image converter tube ME 1201

### *Grid electrode*

In designing an image converter for the purpose described, the avoidance of inertia effects and the furtherance of controllability stood foremost. In the first experiments, carried out with existing standard image converter tubes, these were controlled by pulsing the high anode voltage (usually between 6 and 12 kV). Easier and faster operation is made possible by introducing an extra electrode acting as a grid with a good control characteristic. This has been done in the ME1201 (*fig. 5*). The tube current is biased off by giving the grid electrode a potential of about 60 V below cathode potential, and the tube may be triggered by a high positive voltage pulse, of say 2 or 3 kV, on the grid. During operation the screen (anode) is at a potential of 6 kV with respect to the cathode.

A high grid potential is essential for obtaining at the photocathode surface an electric field intensity ensuring the maximum possible electron emission (elimination of influence of space charge). At the same time, this electric field combined with a constant magnetic field produced by a focusing coil as illustrated in *fig. 6* constitutes the electron-optical system focusing the electrons onto the fluorescent screen. This system, which is characterized by low distortion and good definition of the picture obtained, is similar to that used in the image-iconoscope, a type of television pick-up tube recently described in this Review [6]).

The electron-optical system will be considered in more detail presently, but first a few words should be said regarding the photocathode and the fluorescent screen.

### *Photocathode*

The semi-transparent photocathode is of the caesium-antimony type also employed in the image iconoscope and in other devices, e.g. in photomultipliers. It is noteworthy that photocathodes of this type, when made according to the conventional methods, are found to be

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[6]) P. Schagen, H. Bruining and J. C. Francken, The image iconoscope, a camera tube for television, Philips tech. Rev. **13**, 119-133, Nov. 1951.

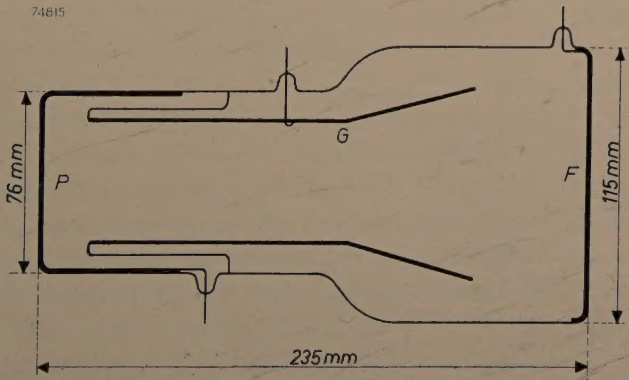

Fig. 5. Cross-section of ME1201 tube. *P* photocathode, *F* fluorescent screen, *G* grid electrode. The latter is formed by a layer of aluminium on a glass tube supported by a reentrant part of the vessel. Near *P* and *F*, the inner wall of the vessel is metallized to connect the photocathode and the metal-backed screen to the cathode and anode lead-in respectively. Part of the vessel contains an outer metallization connected to the cathode in order to minimize the occurrence of internal wall charges that will result in the transmission by the tube of spurious signals.

completely unsuitable for our purpose: when it is tried with such a cathode to photograph the discharge of a fast electronic flash tube as e.g. the LSD2, no picture at all becomes visible on the fluorescent screen of the image converter. This is due to the photocathode not being able to conduct in such a short interval the required number of electrons to the parts which are subjected to an intense illumination and from which the electrons should be emitted (*fig. 7*). In fact, the surface resistance responsible for this failure amounts to several megohms per square [7]) in conventional photocathodes. A special caesium-antimony cathode was developed, having a surface resistance of only a few hundreds of ohms per square. This value has proved to be small enough, as now the actual limitation in decreasing exposure times in most cases is no longer imposed by the cathode. In reducing the surface resistance, part of the overall sensitivity of this type of photocathode had to be sacrificed; it amounts to 20 μA/lumen for incandescent light (from tungsten at 2700 °K), as against values of 45 or even 60 μA/lumen that have been obtained with conventional photocathodes.

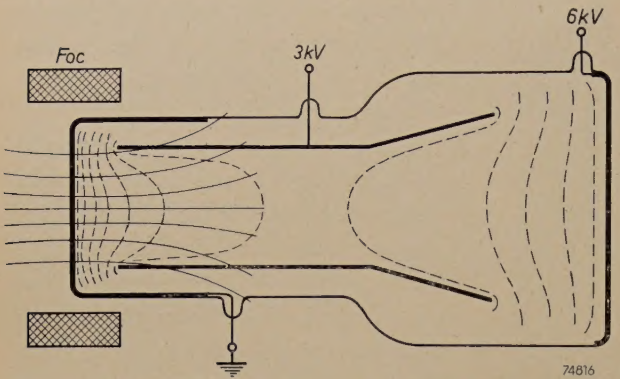

Fig. 6. Electron-optical system for the ME1201. The focusing coil *Foc* produces a magnetic field whose lines of force are indicated by thin lines. The grid and anode potentials establish an electric field whose equipotential surfaces are indicated by dotted lines. The joint action of these two fields on the electrons emitted from the photocathode produces an image of the photocathode on the fluorescent screen.

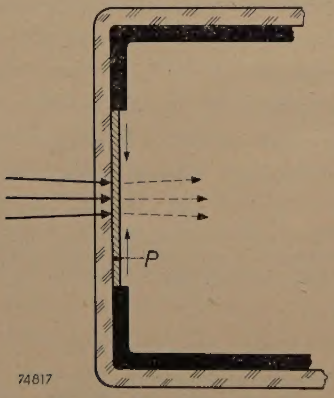

Fig. 7. The useful part of the photocathode *P* must be very thin (semi-transparent) in order to transmit the light to the inner surface, where the electrons must leave the metal (dotted arrows). In supplementing these electrons from the surrounding ring-shaped metal layer, the photocathode must carry a current parallel to its surface (thin arrows). For adequate negative densities produced with an exposure time of $10^{-7}$ sec, peak currents of tens of milliamperes must be drawn from the cathode.

The spectral response curve of the low-resistance caesium-antimony photocathode is given in *fig. 8*. It resembles the spectral response curve of the eye, except for a shift to the blue. This response curve is ideally suited for photographing objects illuminated by an LSD2 flash tube, as may be seen from the spectral energy distribution curve of the latter (*b* in fig. 8). For other light sources other photocathodes can be used, for instance the well known Ag-O-Cs cathode which is sensitive to infra-red radiation, or a combination of one of these cathodes with a phosphor sensitive to ultra-violet radiation or to X-rays.

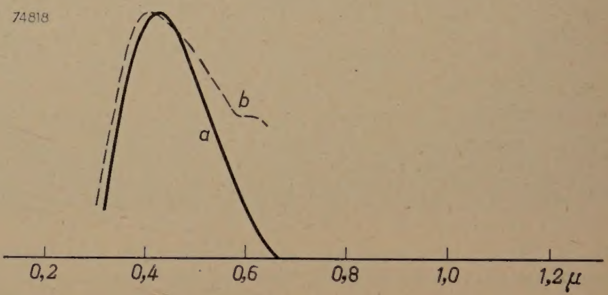

Fig. 8. *a*) Spectral response curve of low-resistance caesium-antimony photocathode (on a relative scale.)
*b*) Spectral energy distribution curve of LSD2 electronic flash tube (on a relative scale).

### *Fluorescent screen*

The fluorescent screen material in most cases adopted for the ME1201 is zinc sulphide. Alternatively willemite ($Zn_2SiO_4$) may be used. The former radiates about 3 times more energy per second than the latter, when excited by 6 kV electrons. Moreover, the spectral distribution curve of the radiation emitted by a zinc sulphide screen is better adapted to the response curve of the high sensitivity photographic emulsions that must be used for this work (cf. *fig. 9*). Both effects combine to make zinc sulphide 5 times more efficient photographically than willemite, so that the former material is preferable for general purpose use. Another difference between the two materials is the slower decay of the luminescence usually exhibited by a willemite screen.

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[7]) The surface resistance of a layer of given resistivity and thickness only depends on the configuration of the conducting surface, not on the dimensions.

Curiously enough, this might be considered as an advantage for single shot photography. No blurring is introduced by the fact that light is being emitted from the screen for a long time after the electrons conveying the image have been cut off, and in fact this light is beneficial to the density of the negative obtained, as the sensitivity of any emulsion will decrease at very short exposure times (reciprocity failure or Schwarzschild effect). For multiple shots with a high repetition rate, however, a short afterglow duration is essential in order to avoid that each frame should suffer interference from the preceding ones. With a zinc sulphide phosphor containing nickel as a "killing" agent, decay times as short as 10 μsec have been obtained (drop from peak intensity to 25%).

The efficiency of the fluorescent screen is greatly increased by an aluminium backing, formed in the usual way [8]) and reflecting most part of the light that is emitted backwards and would otherwise be lost. At the same time, this portion of the emitted light is prevented from hitting the photocathode and producing extra electrons that would cause severe fogging of the fluorescent image. In our specific case the metal backing layer fulfils two more functions of equal importance. Firstly the caesium vapour which is present in the tube during the formation of the photocathode is prevented from contaminating the phosphor. Secondly the light transmitted by the photocathode — and which in this application of the image converter in general has about the same wavelengths as the light emitted from the screen! — is not allowed to illuminate the screen. Evidently, this is a *conditio sine qua non*, as such an illumination would be liable completely to spoil the pictures to be taken. Even with the metal backing the avoidance of this effect is by no means easy, as the metal layer must be thin enough for transmitting practically all the 6 kV-electrons. Considerable care is needed to produce such aluminium layers not containing sections too transparent to light.

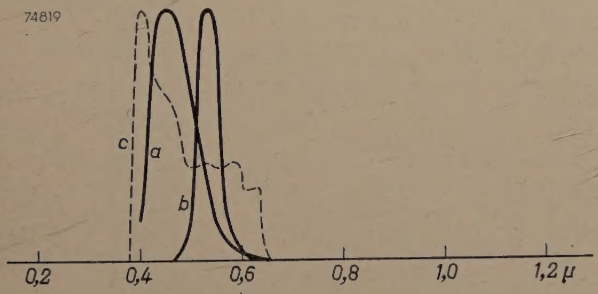

Fig. 9. *a*) Spectral energy distribution of fluorescence of a typical zinc sulphide screen.
*b*) The same for a willemite screen.
*c*) Spectral sensitivity of a fast panchromatic emulsion.

Both willemite and zinc sulphide can be made to form very fine grained screens, capable of resolving 50 lines/mm [9]). This resolution is high enough to ensure that in most cases the performance of the image converter tube will not depend on the resolving power of the fluorescent screen. The following paragraphs will make this clear.

### *Electron-optical system*

It was already mentioned that the electron-optical system comprises a magnetic field produced by a coil around the cathode part of the tube and an electric field produced by the grid and anode potentials. In fact, the electric field existing between the grid and the anode should be considered as a separate, electrostatic lens. For the moment, however, this can be neglected. The electron-optical system produces at the fluorescent screen an image of the photocathode with a magnification of about 4 diameters and with a rotation of about 20°. Such a rotation, inherent in all magnetic electron lenses, is of no consequence whatsoever in our case. An essential property of the combined magnetic and electric electron-optical system is that a nearly flat image is obtained from a flat photocathode, thus facilitating the manufacture of the tube and enabling the use of normal photographic camera optics at both ends in the set-up.

The distortion of the picture is found to depend greatly on the absolute values of both the magnetic and electric field intensities. Among different types, pincushion distortion and S-bend distortion are most prominent, the latter being typical of all magnetic electron-optical systems (cf. [6])). Both effects are proportional to $d^3$, $d$ being the diameter of the useful part of the photocathode. The important influence of the electric field intensity at the cathode is demonstrated by the series of pictures in *fig. 10*, taken with different values of the grid potential $V_g$. It is seen that with the cathode diameter $d \approx 2.5$ cm used in the ME1201 the S-bend and other distortions become negligible at $V_g >$ about 2.5 kV.

It should be noted that for each value of $V_g$ (and $V_a$ if this were changed), the focus coil current must be adjusted anew so as to bring the image in sharp focus on the fluorescent screen. It is interesting to

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[8]) See for instance: Philips tech. Rev. **10**, 101-102, 1948.

[9]) I.e. a pattern containing 25 black and 25 white lines per mm. It is usual in television technique to designate this as 50 lines/mm, whereas according to conventional photographic terminology this would be called 25 lines/mm.

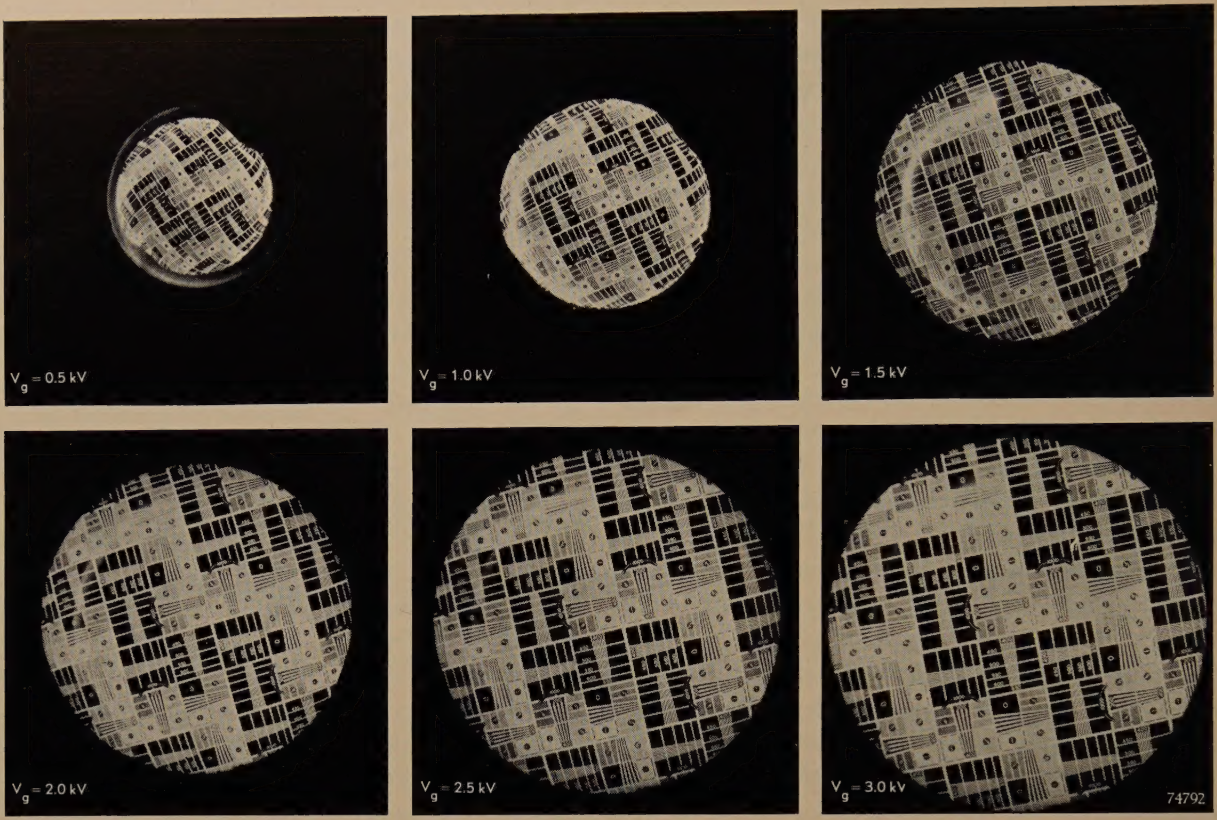

Fig. 10. Static pictures recorded from the screen of the ME1201, at an anode voltage of 6 kV and different grid voltages $V_g$. At low grid voltages the image reveals S-bend and other distortions.

note that for each $V_g$ value this can be effected with several distinct current values, the successive images increasing in size with increasing current. The image obtained with the lowest current value ("first focus") suffers from considerable distortion. The pictures in fig. 10 were obtained with the second or third focus, which is also normally used and for which the above-mentioned values of magnification and rotation are valid.

The appearance of several foci can be understood from the mathematical treatment of the image formation in an image iconoscope, given by Francken and Dorrestein [10]). Their results can be summarized for our purposes as follows. The paths of all the electrons emitted from one point of the photocathode with different radial and/or tangential velocities intersect again in a point at a distance $z_0$ from the cathode. The value $z_0$ is found to be the same for electron pencils emanating from all cathode points, i.e. an image of the cathode is formed at a distance $z_0$. In order to compute $z_0$, the electron pencil originating in the center of the photocathode, whose rays will intersect on the axis of the system, is examined. Apart from a rotation, which does not matter in this case, the path of an electron starting with an initial radial energy $\varepsilon_r$ is given by the function:

$$r = 2\sqrt{\frac{\varepsilon_r}{eF}}\, r_2(z),$$

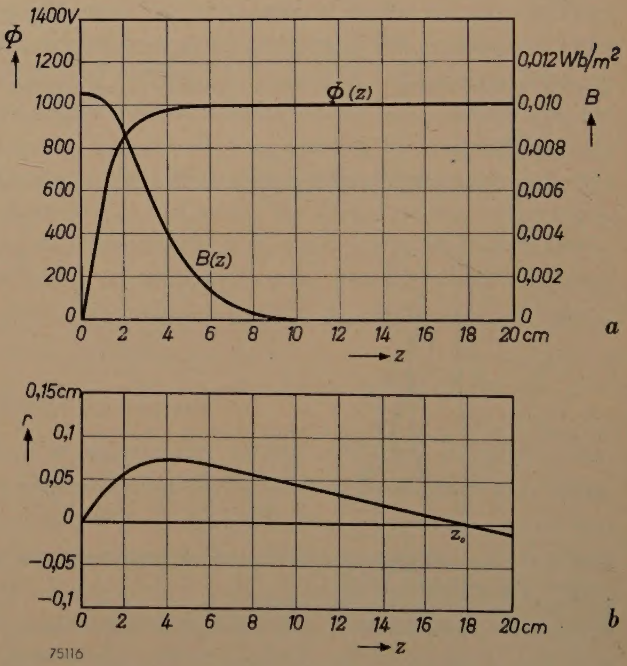

Fig. 11. *a*) Flux density *B* and electric potential $\Phi$ on the axis (*z*-direction) of an image iconoscope as described in the articles quoted in [6]) and [10]). The field configuration is similar to that in the electron-optical system of the image converter ME1201.
*b*) Path of an electron emitted from the center of the photocathode, as calculated by Francken and Dorrestein [10]) for the magnetic and electric fields given in (*a*). The initial radial energy of this electron is $\varepsilon_r$ = 0.65 V; for other values of $\varepsilon_r$, the radial displacements $r(z)$ are proportional, so that all the electrons emitted from the centre of the photocathode intersect the axis at $z_0$.

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[10]) J. C. Francken and R. Dorrestein, Paraxial image formation in the "magnetic" image iconoscope, Philips Research Reports **6**, 323-346, 1951 (No. 5).

$r$ being the radial displacement from the axis, $F$ the electric field intensity at the center of the photocathode, $e$ the elementary charge and $r_2(z)$ a function dependent on the shape of the magnetic and the electric field. The path $r(z)$, calculated for a magnetic and an electric field typical of the image iconoscope and for a practical value of the initial electron energy $\varepsilon_r$, is plotted in *fig. 11.* If the zero point $z_0$ of $r_2(z)$ occurs within the magnetic field region and if this field is intense enough, the rays may be turned back towards the axis again, and more zero points of $r_2(z)$ may occur, i.e. more than one image may be obtained. This evidently is our case. When increasing the magnetic field intensity, the curvature of the rays $r(z)$ is increased, all the images are drawn nearer to the photocathode and the images are successively made to coincide with the fixed plane of the fluorescent screen.

It is further shown by Francken and Dorrestein (formula 62d of their article) that the image magnification is inversely proportional to the final slope of $r_2(z)$, at the respective focus ($z_0$, ...). This slope will be smaller for higher focus order, even when the focus distance in each case is made equal to the tube length. In this way the increased image size of successive foci is understood.

The definition of the electron-optical image obtained at the fluorescent screen is greatly affected by fluctuations of the magnetic and electric field intensities. Careful stabilization of the grid and anode voltages as well as of the coil current is necessary if the highest definition, matching the resolving power of the fluorescent screen, is required. *Fig. 12*, a reproduction of a static picture produced at the screen, may serve to demonstrate the optimum resolution of which the electronoptical system is capable. When microsecond photographs are being made, a lack of stability will not of course exert its influence within the very short exposure period. As the focusing, however, is effected by producing a static picture of the object prior to the actual exposure and by adjusting the anode and grid voltages and the coil current so that this picture when examined visually is sharpest, a change of the adjusted values in the time interval between the

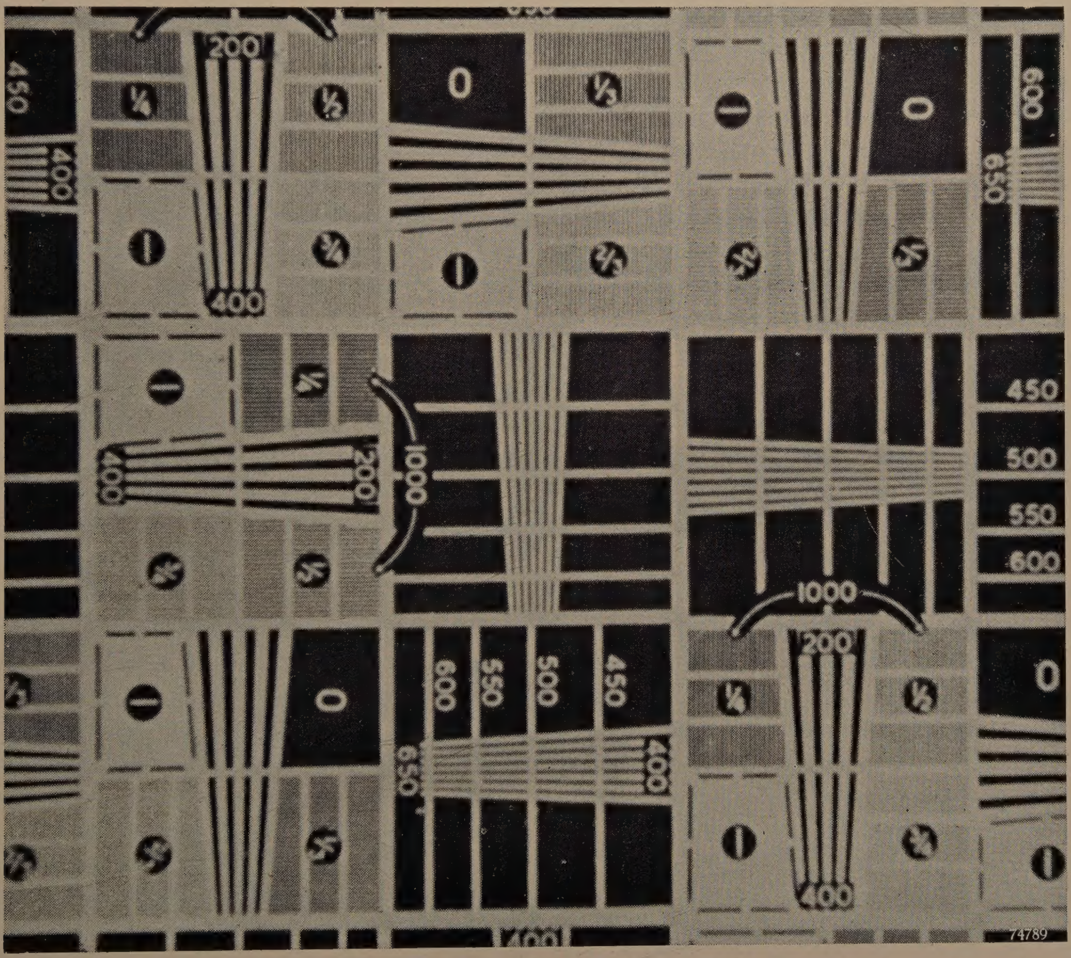

Fig. 12. Static picture of test pattern recorded from central part of the screen of the ME1201 image converter, in order to demonstrate the optimum obtainable resolution. About 3 × size of image on screen. The influence of coil current and grid and anode voltage fluctuations was eliminated and the picture was recorded on a very fine-grained emulsion. The resolution is about 50 lines/mm.

adjustment and the exposure may result in a loss of focus. The stabilization of the grid voltage is especially critical, as this will affect both the electron-optical system formed by the combined magnetic and electric fields and (to a lesser extent) the electron lens formed by the electric field between grid and anode; the focus shift caused by a grid voltage variation unfortunately is for both lenses in the same direction. However, with a ripple of about 1% in the grid and anode voltage and in the coil current, 20 lines per mm will be resolved at the fluorescent screen. For many practical purposes this will do, as it corresponds to about 2000 lines in the entire picture, with a fluorescent screen image diameter of about 10 cm. The total number of lines (the "information" conveyed) then is 4 or 5 times that of a television picture according to American or British standards.

It has been stated that the electron-optical system with the normally used focusing coil provides a linear magnification of about 4×. As nearly all the electrons emitted by the useful part of the photocathode are concentrated in the image at the fluorescent screen, the screen brightness will be proportional to the inverse square of the magnification. Therefore, in cases where a shortage of light is experienced, lower magnification or even a demagnification may be desirable (cf. the article quoted in [3]). With the aid of other focusing coils other magnifications, varying from 7 × down to 2×, are possible. For even lower magnifications, other tubes have been designed. With one of our latest tubes, which is not available as yet for regular distribution, a demagnification by a factor $^3/_4$ is obtainable, resulting in a gain of about 16× in brightness of the fluorescent image as compared with the ME1201 in normal use, all other things being equal. It should, however, be pointed out that in cases where the struggle for sufficient light does not set the scene, a certain electron-optical magnification as used in the ME1201 offers important advantages [11]). A lower magnification will almost inevitably mean a smaller image at the fluorescent screen, as it is very difficult to increase the useful photocathode area without introducing an excessive distortion in the electron-optical reproduction, to say nothing of the difficulty of avoiding local sensitivity variations in large photocathodes. A smaller screen image size, however, is bound to bring the graininess of the screen into greater prominence, affecting the overall resolving power of the method. This point of view is especially important in a procedure for taking multiple shots that will be described at the end of this article. The loss in resolution may partly be restored in the actual recording process. In fact, as long as the shortage of light is predominant, very sensitive emulsions must be used in recording the picture from the screen and the resolution will be limited by the film grain. If the brightness of the screen image is substantially increased, finer grained emulsions of lower sensitivity may be used.

### Triggering and synchronizing circuit for single shots

The general components of the circuit used for taking single shots with the ME1201 have already been indicated in fig. 3. In the "waiting" condition of the image converter the cathode potential is made 3.1 kV, the grid potential 3 kV. No electrons can leave the photocathode in this condition. The tube is triggered (the "shutter is opened") by bringing the cathode potential down to zero by means of a switch $S_1$. It is biased off again (the "shutter is closed") by bringing the grid down to a slightly negative potential, e.g. — 100 V, by means of a second switch $S_2$. The switches $S_1$ and $S_2$, which must be operated in rapid succession, to define the interval of the exposure, consist of hydrogen thyratrons which can be triggered at an accurately fixed time by a comparatively low grid voltage pulse.

The grid-cathode system of the image converter tube possesses a finite capacitance, amounting to appr. 100 pF, that must be charged and discharged in the successive pulsing of cathode and grid potentials. These charging and discharging processes, involving the transport of an electric charge of about $3000 \times 100 \times 10^{-12} = 3 \times 10^{-7}$ coulomb, will take some time, depending on the maximum current the thyratron tubes are able to carry without damage to their cathodes, on the building-up time of this current and on the resistance occurring in the charging circuit. The charging and discharging periods should be made as short as possible, as during these periods electrons are passed from the photocathode to the fluorescent screen. It is true that during these periods the proper focusing conditions are not established, so that these electrons will not introduce a blurring due to the rapid change or movement of the object to be photographed. They may, however, contribute a background detracting from the picture quality.

A direct current of 30 A would be required to deliver the above-mentioned electric charge in $10^{-8}$ seconds. The 3C45 thyratron (of the range of Mullard thyratrons) used in this application is rated at a current of 35 A. However, because of the building-up time of the maximum current and because the current intensity gradually decreases when the charging of the image converter capacitance is nearing its completion, the actual duration of the charging period will be appreciably longer than $10^{-8}$ sec. These are, then, the limiting factors in diminishing the exposure time. The actual limit, i.e. the minimum exposure time providing pictures in which the background effect is not yet excessive, can only be determined from experience. With the present technique a minimum exposure time of 3 to 4 × $10^{-8}$ seconds has proved feasible. This result is partly due to the fact that the maximum current of the 3C45 thyratron can considerably be increased beyond the rated limit for single shots or for low repetition rates.

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[11]) Cf. also: J. S. Courtney-Pratt, Image converter tubes and their application to high speed photography, Photographic J. **92B**, 137-148, 1952 (Sept.-Oct.).

After an exposure has been made, the anode voltages of both thyratrons will gradually return to a high value, both tubes become non-conducting by deionization, and the cathode and grid of the image converter are charged again to the potentials of 3.1 and 3 kV respectively of the "waiting" condition. During this recovery the image converter must constantly remain biased off; this is ensured by arranging the cathode circuit to deionize before and to charge up more quickly than the grid circuit.

The circuit developed for single shots is shown diagrammatically in *fig. 13*; a photograph of the

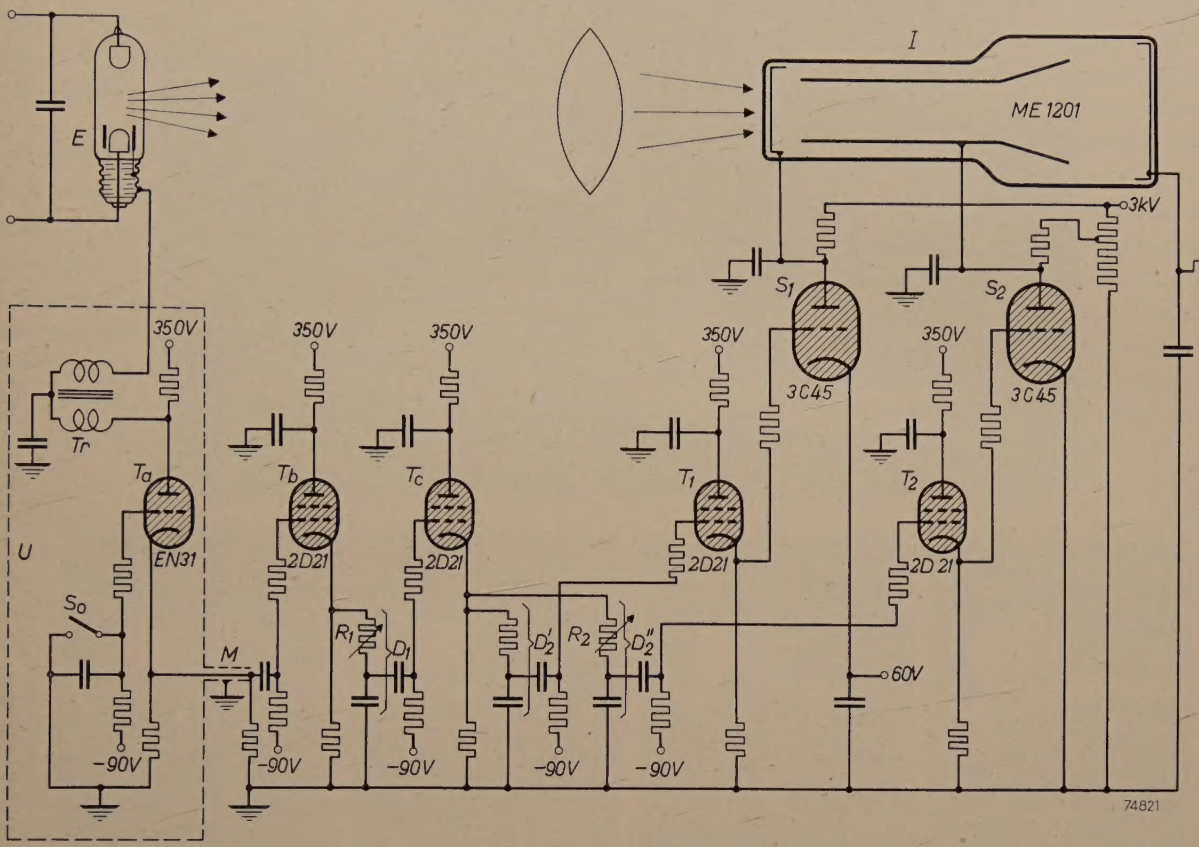


Fig. 13. Circuit for the taking of single shots with the ME1201 image converter tube. A few non-essential details have been omitted. The sequential pulsing of cathode and grid electrode of the image converter *I* is effected by the 3C45 hydrogen thyratrons $S_1$ en $S_2$ respectively (cf. also fig. 3). These are rated at 35 A and can stand a 3 kV anode voltage without breaking down at zero grid voltage. They are fired by the smaller 2D21 thyratrons $T_1$ and $T_2$, which receive their triggering grid pulse voltages from a thyratron $T_c$ via two delay circuits $D_2'$ and $D_2''$ respectively. $D_2'$ gives a fixed delay, whereas the delay caused by $D_2''$ can be finely controlled by means of the variable resistor $R_2$, so that the switch $S_2$ pulsing the image converter grid can be actuated before or up to 2 μsec after $S_1$ pulsing the image converter cathode. Thus, $R_2$ sets the exposure time. The thyratron $T_c$ in turn is triggered when the mechanical switch $S_0$ is closed, via thyratrons $T_a$ and $T_b$, and another delay circuit $D_1$, adjustable by means of the variable resistor $R_1$ to give a delay of up to 100 μsec. This serves to synchronize the firing of the image converter with the peak of the light emission of an electron flash tube *E*, which is triggered by $S_0$ via thyratron $T_a$ and a transformer *Tr*. (The thyratron $T_b$ is interposed because the pulse supplied by the conventional electron flash tube triggering unit *U* is not suitable for triggering $T_c$. The coaxial cable *M* connecting *U* to $T_b$ must correctly terminated to avoid the occurrence of reflected signals which may fire $T_b$.)

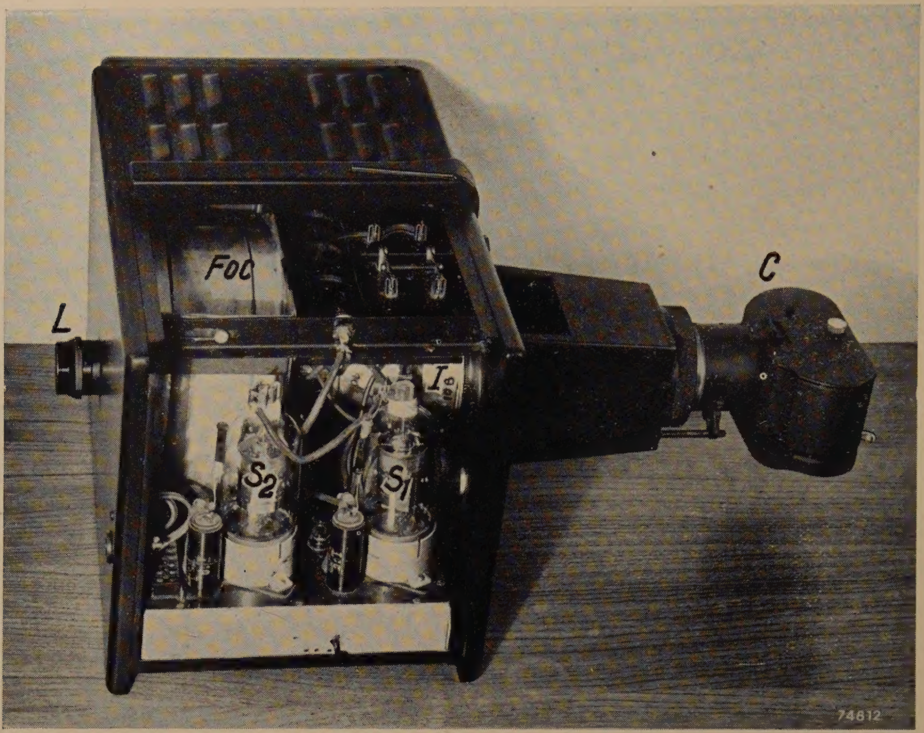
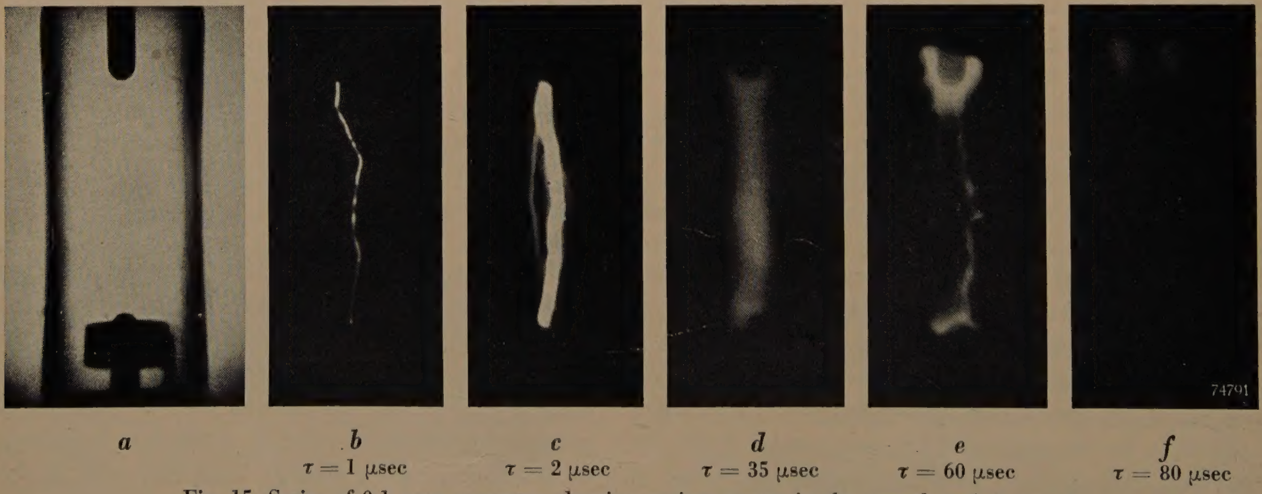


Fig. 14. Laboratory camera unit for high speed single shot photography. The image converter tube $I$ is mounted in a transverse position in the cabinet. To the left the lens $L$ producing an image of the subject on the photocathode is seen. The recording camera $C$ projects to the right. In the near part of the cabinet the two hydrogen thyratrons $S_1$ and $S_2$ are seen.

apparatus comprising the circuit components and the image converter tube ME1201 is represented in *fig. 14*. From fig. 13 it is seen that the two large hydrogen thyratrons 3C45 ($S_1$ and $S_2$) are triggered by two smaller thyratrons, of the 2D21 type ($T_1$ and $T_2$). The firing of these two small thyratrons is effected by yet another 2D21 thyratron ($T_c$) via a pair of integrating circuits ($D_2'$ and $D_2''$), introducing different delay times for $T_1$ and $T_2$. The time interval between the firing of $T_1$ and $T_2$ is equal to the nominal exposure length and can be adjusted by means of the variable resistor $R_2$ to any value up to 2 μsec. Finally, the firing of the thyratron $T_c$ that starts the exposure is connected with the firing of the small thyratrons $T_a$ and $T_b$ in order to enable the exposure to be accurately synchronized with the peak of the luminous output of an electronic flash tube illuminating the object to be photographed. The legend of the figure gives some additional information as to this synchronization. The accuracy



*a*

*b* $\tau = 1$ μsec

*c* $\tau = 2$ μsec

*d* $\tau = 35$ μsec

*e* $\tau = 60$ μsec

*f* $\tau = 80$ μsec

Fig. 15. Series of 0,1 μsec exposures showing various stages in the growth and decay of the discharge in an LSD2 flash tube. *a*) is a static picture of the electrodes, *b*)-*f*) are made at increasing times $\tau$ after triggering.

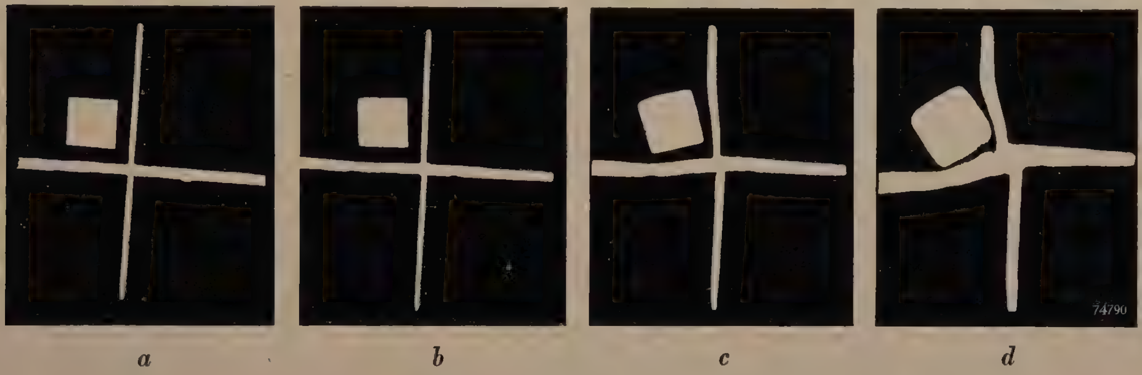
of the synchronization is aptly demonstrated by *fig. 15*, showing a number of $10^{-7}$ sec exposures made of the discharge of the LSD2 electronic flash tube itself. The overall "jitter" is so slight that pictures of the type of fig. 15*b* can be reproduced at will in successive flashes of the tube. Incidentally, it will be clear from the pictures that this method of high speed photography is a powerful tool in the investigation of the mechanism of discharge of these electron flash tubes and the like.

### Special effects in very short exposures

Several peculiarities in the behaviour of the "image converter shutter" at very short exposures have already been mentioned, e.g. the influence of the surface resistance of the photocathode delaying the supplementing of electrons to heavily emitting cathode parts; the gradual increase and decrease of the grid voltage owing to the limitations in the current supplied by the switching thyratrons; the possible transmission of light through some areas of the backing layer of the fluorescent screen during the waiting period. Two other and rather important effects should be mentioned here.

The first of these is related to the presence on the grid of minute quantities of photosensitive material deposited there during the manufacturing of the tube. If the illumination or the light emission of the object to be photographed persists during the waiting period, the light transmitted by the semitransparent photocathode when striking the grid will liberate electrons from it. These electrons are accelerated and crudely focused by the electric field between grid and fluorescent screen, thus giving rise to a ring of light within the picture area (cf. fig. 10). As this effect, feeble in itself, may be integrated over a relatively long time, it may in some cases be liable to ruin the record. The only general remedy for these cases will be a lowering of the anode voltage or a raising of the grid voltage.

The second effect to be described is probably again due to the surface resistance of the photocathode mentioned previously. It consists of a local distortion of the picture at places where very bright high lights occur. The effect is illustrated by *fig. 16*. Although its explanation is not yet quite certain, it seems to us very probable that it is due to the electrons emitted from the highly illuminated portions of the photocathode not being supplemented at a sufficient rate to prevent these portions from charging up to near grid potential during the very short exposure interval. Such a transient local charge may produce transverse fields at the surface of the photocathode which will exert a powerful influence on low velocity electrons emitted by other parts of the photocathode.



*a* *b* *c* *d*

Fig. 16. Distortion due to locally overloading the photocathode of the image converter by very high illumination. A simple transparency, as shown by the static picture (*a*) was imaged onto the photocathode with different objective apertures, and exposures of 0.1 µsec were made. In (*c*) and (*d*) the illumination is 2 × and 4 × times that in (*b*), respectively.

The distortion effect is rather important as it sets another limit to the minimum exposure time. From the photographic point of view this is rather strange a limit because it means that the minimum exposure depends on the brightness range rather than on the brightness of the object to be photographed. In fact, it was found in a specific case using an average image converter tube, that distortion occurred if a high light density greater than 0.5 above fog was recorded in an exposure of 0.1 µsec. The practical conclusion is that for the shortest exposures the suitability of emulsions and developers should not be judged by their properties for normal use, but chiefly by their ability to bring out details at very low densities of the negative. Fortunately this ability is highest with the fast panchromatic emulsions.

### Use of the image converter tube for multiple shots and scanning

If very short exposures are necessary to avoid blurring by changes or movements of the object, a cinematographic record, i.e. a continuous representation of these changes of movements, obviously will require a series of pictures with very short intervals, comparable to the exposure time. Thus frame rates between $10^6$ and $10^7$ per second must be aimed at. The recovery period that was mentioned in the description of the circuit used for single shots, and that is necessary inter alia for the deionization of the switching thyratrons, will prohibit any increase of the repetition rate beyond about 20 kc/s. Much higher repetition rates are made possible by the application of radar modulator techniques. In that case a limit to the shortness of the intervals between successive pictures will rather be set by the speed with which the film can be transported and by the decay of the fluorescent screen image. By the use of high-speed drums and of a suitable very fast decaying screen, film rates of say 200 000 frames/sec should be made possible. When making exposures of $10^{-7}$ sec, however, the intervals in this case would be still 50 times the exposure time. Thus, a real cinematographic record of the very fast events that can be stopped by the image converter method cannot be achieved in this way.

There are three different methods by which the difficulty of high frame rates can be overcome. The first one is the well-known stroboscopic method, which can be applied in the case of periodic events: successive phases of the event are not picked out from one single period but from successive periods or even hopping over a number of periods. Demonstrations of this use of the image converter have been given showing as an example the movement of stress waves in a transparent medium. A single shot circuit was used, operating the image converter shutter at 20 kc/s with 1 μsec exposures. The screen was observed visually; for this application a fast decay phosphor is not necessary, as long as the afterglow does not exceed the apperception time of the eye.

The second method consists in producing an image covering only part of the fluorescent screen, this small image being successively deflected to different positions on the screen. To this end, two pairs of deflecting coils, for horizontal and vertical deflections respectively, are fitted around the image converter tube. To each pair of deflection coils a current having one of a number of discrete intensity values can be fed, these values being alternated in such a way, that the necessary combined deflections for the different image positions are obtained. At low speeds, e.g. for deflection intervals of the order of 1/1000 second, a mechanical commutator can be used for alternating the deflection current values, and this will be quite a useful method for the analysis of many problems in research and industry. The same principle, however, using electronic means for alternating the deflection currents should allow of image intervals and image stopping times of the $10^{-7}$ sec range attained with single exposures. It will probably be possible with future types of tubes having larger screens to give the image up to 25 different positions on the screen, enabling a sequence of 25 pictures to be taken by a single shot (9 positions have been obtained already). A combination of the deflection method with the previously mentioned high speed cine-camera techniques then will give an effective multiplication of up to 25 in frame rates.

Even more powerful means for recording extremely rapid changes are available if the information desired can be conveyed by a picture in only one dimension, preferably a straight line. Such is the case with e.g. the propagation of an explosion, the starting of a discharge etc. The idea of getting separate distinct images that was adhered to in the preceding method can then be dropped and a continuous deflection of the (linear) image is possible. This continuous deflection or scanning is effected by one pair of deflection coils similar to those mentioned above. It should be mentioned that both in this method and in the preceding one, no separate pulsed grid voltage is required, and in fact a simpler converter tube (the ME1200) having no grid electrode may be used in this case. Using a sinusoidal deflection coil current, very high deflection rates can be obtained. Useful exposures were obtained by Courtney-Pratt ¹¹) at deflection rates of up to 300 000 metres per second; this is a writing speed about a thousand times higher than that achieved mechanically with the best drum cameras. The "time resolution" obtained with these deflection rates is of the order of $10^{-9}$ sec. In fact, Courtney-Pratt was able in this way to measure with reasonable accuracy the time difference in the arrival at the photocathode of two light pulses that had travelled along paths differing by a few feet in length.

### Comparison with other methods of high speed photography

To conclude this article a few remarks should be made on the relative merits of the image converter technique when compared with other methods permitting exposure times of the same order ¹²). Among

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¹²) A survey has been given by J. M. Meek and R. C. Turnock, Electro-optical shutters as applied to the study of electrical discharges, Photographic J. **92B**, 161-166, 9152 (Sept.-Oct.)

such methods the Kerr cell shutter is the oldest and still flourishing. When operated with electronically generated voltage pulses it permits of minimum exposure times of the same order as the image converter, i.e. $10^{-7}$ sec and even less, but its inherent disadvantage is the limited sensitivity. This is due partly to the light absorption in the shutter proper (the maximum transmission being only 5% of the incident light) and partly to the limited angle of the light beam transmitted. With the image converter on the contrary, the "transmission" can be well over 100% in the shutter proper, owing to the energy being imparted to the electrons by the high anode voltage, and the complete range of wide aperture, wide angle lenses can be used. In fact, the deficiency of light that was mentioned repeatedly in the preceding paragraphs is not the fault of the image converter which is a very powerful photographic device indeed, the effective aperture of the whole apparatus being higher than that of any Schmidt camera (cf. [11]), but it is the fault of the very shortness of the exposure times. As will be well-known to every amateur photographer, when taking an average near-by subject in bright sunshine, an exposure of the order of 1/500 sec at *f*:8 can give a reasonable negative with a fast emulsion. Accepting a value of *f*:1.4 for the effective aperture of the set-up with image converter, and ignoring the Schwarzschild effect, the same subject brightness would require an exposure of about $2 \times 10^{-4}$ sec. Thus for a $10^{-7}$ exposure an illumination several thousand times that in bright sunshine will be necessary.

Other methods devised for recording pictures with very short exposures make use of conventional television techniques. An image of the object to be photographed is projected onto the photocathode of a television pick-up tube, the emission of the photocathode being controlled to occur only during the desired exposure period and an optical image being displayed on a monitor cathode-ray tube. Experiments carried out by Prime and Turnock indicated that, when using a conventional iconoscope as a pick-up tube, the suppression of the image in the waiting condition is rather incomplete and, therefore, the picture quality is spoilt by spurious image formation. Better pictures and higher sensitivity might be obtainable by using an image orthicon as proposed by Meek and Turnock [12]), in which case the image section of the orthicon should be controlled. The device is then fundamentally similar to the image converter method described in this article, except for an increase in sensitivity obtained by the amplification section of the orthicon. It will hardly be necessary to point out that the complete set-up for this method will be much more complex and more expensive than that for the image converter method.

**Summary.** As the electron image in an image converter can be suppressed and released without any inertia being involved, this device can be used as a photographic "shutter" for extremely short exposures. Exposure times as short as $3 \times 10^{-8}$ seconds have been obtained with a special image converter tube, the ME1201, developed in the Mullard Laboratories at Salfords (Surrey, England). This image converter tube is provided with a grid electrode, which during operation is given a potential of 3 kV with respect to the cathode, the anode voltage being 6 kV. Moreover the ME 1201 contains a special photocathode having a very low surface resistance, and a fine-grain fluorescent screen with metal backing. The electron-optical system is similar to that of the image iconoscope; it produces an image of the photocathode (useful diameter 2.5 cm) with magnification between 2 × and 7 × (normally about 4 ×) and practically free from distortion. The optimum resolution obtainable in a static picture is 50 lines/mm (25 black + 25 white ones); in practical cases, with a ripple of 1% in the voltages and currents, the resolution can be 20 lines/mm. The minimum exposure time depends i.a. on the rate of charging the grid-to-cathode capacitance of the image converter, on the limited brightness of the subject to be photographed and on the brightness range of the subject; this is explained in the article. Applications of the instrument with exposure times of $10^{-7}$ sec have already been made in a number of investigations of objects changing or moving at a very high rate. The object may be self-luminous (explosions, sparks, discharges) or it may be illuminated by a light source such as the LSD2 electronic flash tube, having a flash duration of 2 µsec (projectiles, high pressure jets, stress waves, etc.). A circuit for controlling the image converter tube and for synchronizing its operation with that of the light source is described in the article. For cinematographic records the possibility of giving the image a number of different positions, e.g. 25, on the fluorescent screen by means of deflection coils might be very useful. With a one-dimensional image, the method of continuous deflection (scanning) has been applied, permitting of writing speeds up to 300 000 m/sec and of a time resolution of $10^{-9}$ sec.

# HIGH-POWER TRANSMITTING VALVES WITH THORIATED TUNGSTEN CATHODES

by E. G. DORGELO.

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*Until recently the cathodes of transmitting valves delivering 100 kW or more have had to be made of pure tungsten, even though more highly emissive materials have been known for some decades. The development of very active getters, however, has made it possible to produce a transmitting valve with a thoriated tungsten cathode. Cooled with water, this valve will deliver an H.F. output of more than 100 kW with a very high efficiency. Although primarily intended for frequencies up to 30 Mc/s, the valve has also yielded good results in an experimental T.V. output stage at a frequency of 68 Mc/s.*

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## Tungsten and thoriated tungsten as cathode materials

In the last ten years an important development has taken place in the design of large transmitting valves. Formerly, valves delivering 100 kW or more were always fitted with a filament of pure tungsten serving as a cathode. The limited emission of such a cathode necessitates a very high anode voltage; for example the TA 18/100 and TA 20/250 which, although out of date, are still in use today, operate at 18 to 20 kV on the anode. The high field strength prevailing in such valves tends to produce the troublesome phenomenon known as Rocky Point effect [1]), which takes the form of discharge between the electrodes at irregular intervals. These discharges not only lead to interruptions in the service, but also to untimely failure of the valve. Efforts have been made to avoid this by designing valves to operate at lower anode voltages, which in turn necessitated the use of cathodes with a higher emission than in valves for the same power with higher anode voltage. With tungsten cathodes this means that the filament consumption runs into some tens of kilowatts.

A better solution is found in the use of a cathode material having a higher specific emission than pure tungsten, for example tungsten containing about 2% of thorium oxide. Thoriated tungsten is employed for the cathodes of many small transmitting valves [2])[3]) and certain X-ray valves [4]), but its use in large valves was formerly precluded by the high sensitivity of the cathodes to "poisoning"; the extremely thin skin of thorium oxide responsible for the emission is attacked by the slightest traces of gas. Many attempts to incorporate this type of cathode in the larger transmitting valves, in which the vacuum cannot meet the same requirements as in smaller valves, have come to grief as a result of this poisoning effect.

The fact that thoriated tungsten has nevertheless been used with success in the valves described in the following paragraphs may be ascribed to the very high activity of the getters employed. No fewer than three different, carefully prepared getters have been introduced into the valve at suitable points, ensuring a very high vacuum and, in consequence, high and constant emission. This is amply borne out by the fact that such valves worked for years without giving any trouble, both in broadcasting transmitters and in H.F. generators for industrial use (heating).

## General description of the valves

The new transmitting valves are manufactured in two types, one water-cooled (TBW 12/100) and one air-cooled (TBL 12/100). The first of these is depicted at the left in *fig. 1*; on the right will be seen an older type of valve, the TA 20/250 [5]). Both the TBW 12/100 and the TA 20/250 deliver a peak output of about 250 kW in telephony transmitters with anode modulation. The new valve is at once

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[1]) J. P. Heyboer, A discharge phenomenon in large transmitter valves, Philips tech. Rev. **6**, 208-214, 1941.

[2]) E. G. Dorgelo, Several technical problems in the development of a new series of transmitter valves, Philips tech. Rev. **6**, 253-258, 1941.

[3]) E. G. Dorgelo, Glass transmitting valves of high efficiency in the 100 Mc/s range, Philips tech. Rev. **10**, 273-281, 1949.

[4]) J. H. van der Tuuk, Philips tech. Rev. **8**, 204 and 205, 1946.

[5]) The type-numbering code is as follows:
Letters. The letter T indicates that the valve is a triode (Q = tetrode, P = pentode). "A" denotes that the cathode is made of pure tungsten, "B" of thoriated tungsten (C = directly heated, E = indirectly heated oxide-coated cathode). The letter W shows that the valve is water-cooled; in the case of the TA 20/250 the W was omitted because it was obvious at the time that such a large valve would be water-cooled. Forced air-cooling is indicated by the letter L.
Figures. The figures preceding the oblique stroke give the direct anode voltage in kV, and those following the stroke (very roughly) the effective output power in kW delivered by the valve continuously without modulation.

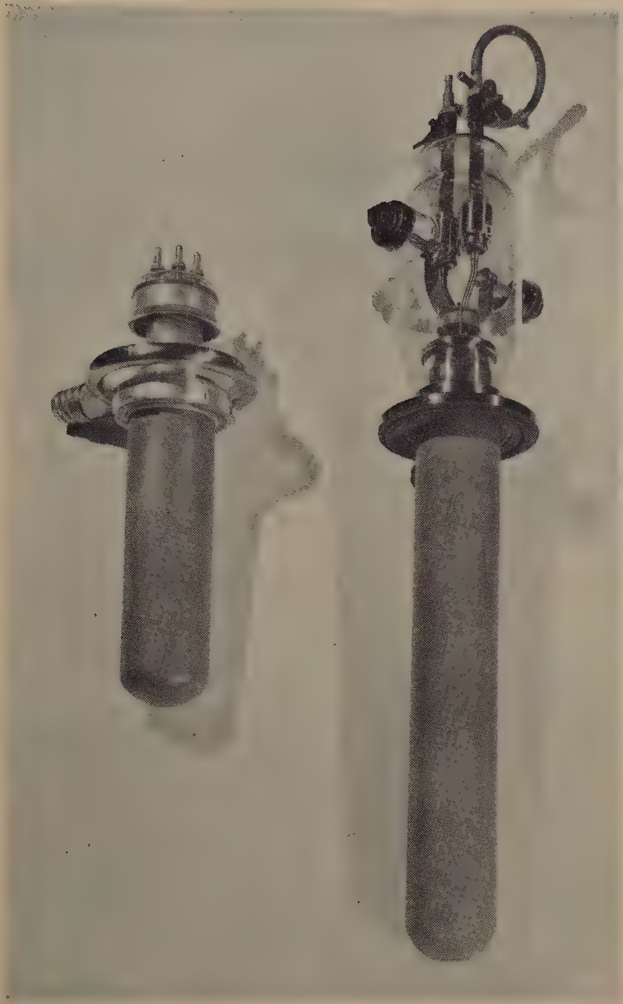

Fig. 1. Left: a modern 100 kW transmitting valve, type TBW 12/100.
Right: Older type of transmitting valve, TA 20/250. When used in a transmitter with anode modulation, they can deliver the same peak output power (250 kW).

remarkable for its smaller dimensions, from which it may be correctly assumed that the losses are smaller and the efficiency higher than in the older type. The type number of the valve reveals the lower anode voltage, viz. 12 in place of 18 or 20 kV.

### *Getters*

*Fig. 2* shows the cathode of the valve, consisting of 12 filaments of thoriated tungsten. Within the circle of filaments there is a molybdenum cylinder *3*, which serves as the support for one of the three getters which takes the form of a layer of zirconium on the wall of the cylinder; radiation from the cathode heats this layer to a temperature of about 800 °C, resulting in a powerful gettering effect, at any rate as regards oxygen and nitrogen [6]). The quantity of zirconium present is sufficient to bind all the gas that the valve would be capable of holding if it were filled with $N_2$ at a pressure of 6 mm Hg or $O_2$ at 10 mm Hg, that is, one million times more than the amount of gas usually remaining after exhausting.

This getter therefore has a very high capacity but, at the same time, it works rather slowly and has to be supplemented by another which works rapidly but need not absorb such large quantities of gas.

This second getter consists of a thin layer of zirconium prepared in a different manner from the first and applied to the grid. It assumes the temperature at which it will take up nitrogen and oxygen only when the driver stage, which causes the grid to dissipate energy, is operating. The greater the amount of power supplied to the grid, the higher the temperature of the grid itself and the more rapid the absorption of gas. The maximum permissible continuous grid dissipation is 2.5 kW, and

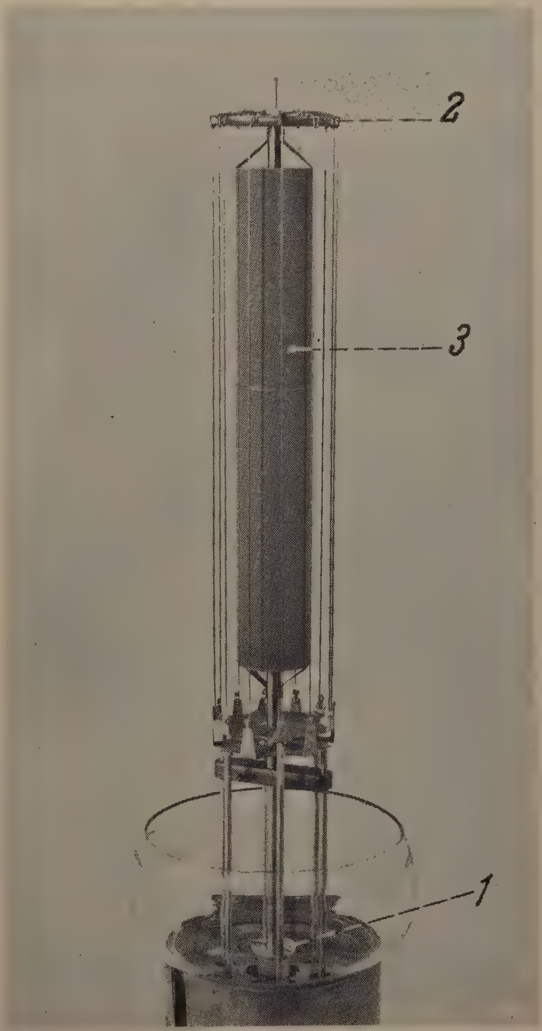

Fig. 2. The cathode of the TBW 12/100 and TBL 12/100 consists of 12 filaments of thoriated tungsten. *1* filament tension spring. *2* springs for absorbing small differences in the filament lengths. A getter *3* is mounted inside the circle of filaments.

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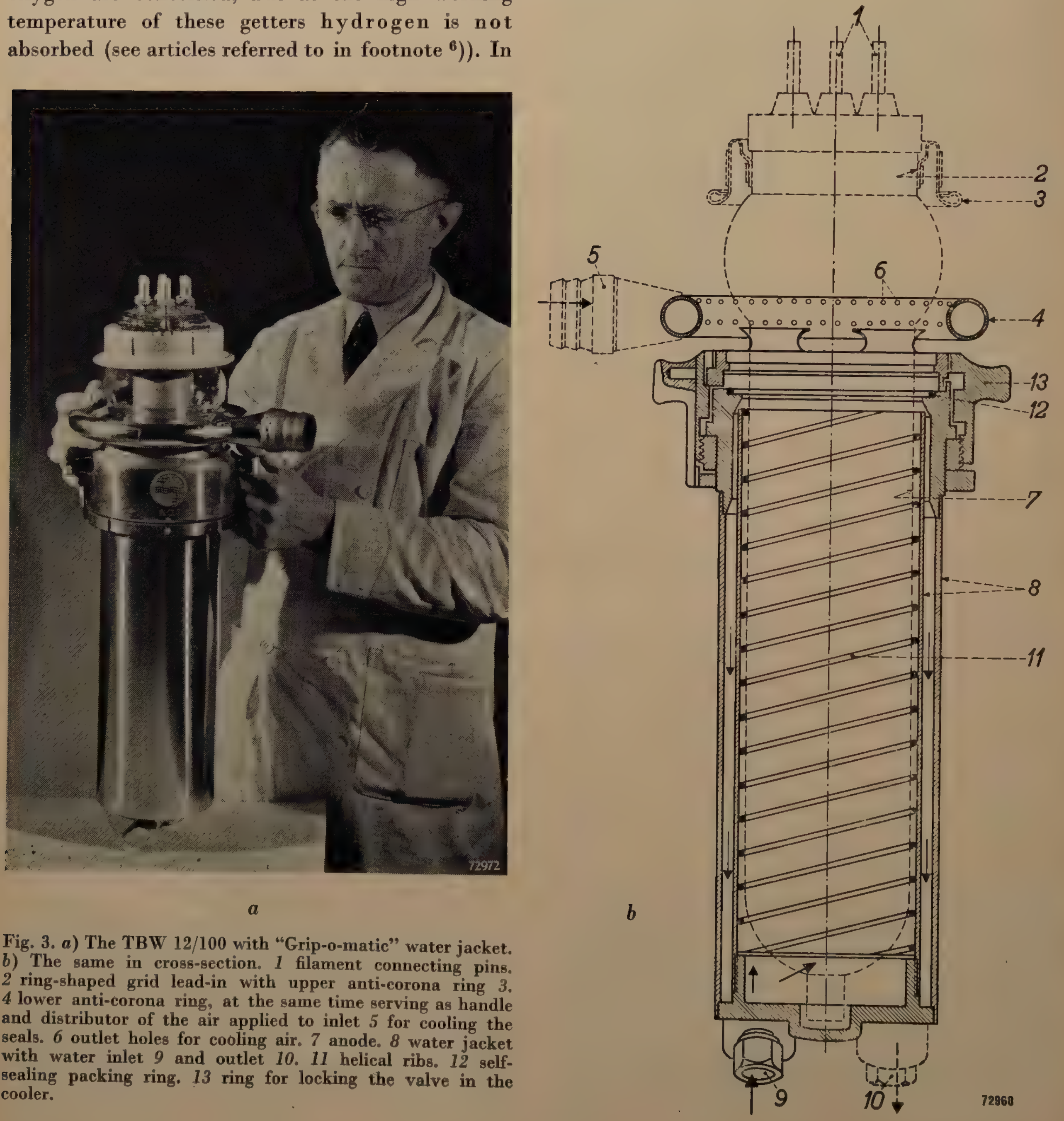
[6]) J. D. Fast, Metals as getters, Philips tech. Rev. 5, 217-221, 1940. See also the article referred to in footnote [2]), p. 255 onwards.

if this limit is exceeded too far, or for too long, there would be a risk of zirconium alloying with the core metal, with a consequent drop in the original high radiation factor of the zirconium layer. The temperature would then rise to a point where grid emission would be intolerably high. However, a special method of preparation of the zirconium powder used ensures that the temperature level at which alloying takes place is increased from roughly 1200 °C to about 1600 °C, which temperature occurs only with a grid dissipation of 5 kW, i.e. twice the maximum rated value.

The two getters described maintain the required high vacuum very effectively as far as nitrogen and oxygen are concerned, but at the high working temperature of these getters hydrogen is not absorbed (see articles referred to in footnote [6]). In order to remove any possible traces of hydrogen in the valve, a third getter is employed, again zirconium, but this time so processed and at such a working temperature that it will readily absorb this gas. This getter becomes active at temperatures below 200 °C and is accordingly placed at the bottom of the anode, in the form of a disc mounted in such a way as to be in close contact with the cooled wall of the anode. The temperature on the inside of the wall is at most 100 °C in the water-cooled valve (max. anode dissipation 100 kW), or 190 °C in the air-cooled type (max. anode dissipation 45 kW). In both cases residual hydrogen is effectively absorbed.



*a*

*b*

Fig. 3. *a*) The TBW 12/100 with "Grip-o-matic" water jacket. *b*) The same in cross-section. *1* filament connecting pins. *2* ring-shaped grid lead-in with upper anti-corona ring *3*. *4* lower anti-corona ring, at the same time serving as handle and distributor of the air applied to inlet *5* for cooling the seals. *6* outlet holes for cooling air. *7* anode. *8* water jacket with water inlet *9* and outlet *10*. *11* helical ribs. *12* self-sealing packing ring. *13* ring for locking the valve in the cooler.

From tests on an experimental valve, so arranged that small quantities of different gases could be introduced into it, it was found that even traces of gas may give rise to breakdown, but that the joint action of the three getters, combined with the relatively low anode potential (10 kV direct voltage with anode modulation) completely eliminates this risk.

### *Details of design*

To ensure high and constant emission, the cathode is subjected during exhausting to the carbonising process usual with this type of cathode; this consists in burning the filament in an atmosphere of hydrocarbon to form tungsten carbide in the wire and thus improve the life of the filament. It is a well-known disadvantage of this carbonising that the wire is thus rendered rather brittle, but any risk of breakage, especially during transport, is considerably reduced by omitting carbonising at the ends of the filaments, where the greatest bending strains occur.

Deformation of the cathode and grid as well as of their supports is avoided by employing tungsten or molybdenum for all these components, both of these metals being highly rigid. The filaments are kept taut by a central spring (*1* in fig. 2) made of a special tungsten alloy which retains its resilience even at elevated temperatures; moreover, this spring is mounted at a point where its temperature will remain as low as possible. Small discrepancies in the lengths of the filaments, which are unavoidable in spite of the greatest care taken in assembly, are equalized by "hairpin" springs *2*.

The total filament current is 196 A, and this is applied through three lead-in pins in parallel which require no artificial cooling, in contrast with the TA 20/250 (fig. 1, top right hand; filament current 420 A), which requires water-cooling.

Connection to the grid is effected by means of a "ring-seal" (fig. 1): the self-inductance of the lead-in is accordingly low, this being an avantage when the valve is operating at very high frequencies. Owing to the large diameter of the ring, resistance to the capacitive grid current is small.

All the lead-in wires are either copper-plated or silver-plated, in order to keep losses as low as possible.

The thick-walled anode is made of O.F.H.C. (oxygen-free high conductivity) copper; in the air-cooled type the anode is provided with cooling fins, to which reference is made later.

As already mentioned, the risk of breakdown within the valve is excluded, but the fact has to be taken into account that breakdown can also occur outside the valve, through the atmosphere, in particular between the ring-seal of the grid and the upper rim of the anode. Breakdown of this kind may destroy the glass seals and, in order to avoid this possibility, each of the components concerned is provided with an anti-corona ring (*fig. 3*). The electric field between these rings is such that any flashover will take place more readily between these rings, where it can do no harm, than between metal parts that are sealed to the glass. The lower anti-corona ring simultaneously serves as a handle when the valve is taken from, or inserted in the cooler, thus avoiding strain on the glass seals.

### Electrical data

It has already been said above that, since the saturation current is very high, the valve is able to operate at a fairly low anode voltage, but another condition has also to be met in order to achieve this, viz. low internal resistance.

In general, low internal resistance is ensured by using a cathode having a large effective area, in which respect a large number of filaments of thin wire is better than a small number of thick filaments; however, physical strength determines the extent to which this sub-division of the cathode can be carried out.

The smallest possible spacing of anode and grid, and especially of grid and cathode, also makes for a low internal resistance, but here again considerations of a physical nature impose a limit, for, amongst other things, the valve must be able to withstand transportation over long distances.

Another consequence of increased cathode area combined with reduced inter-electrode spacing is that the capacitances are increased; this results in a drop in the maximum frequency to which the valve can be tuned, and a reduction in the band-width for a given power gain.

It follows, therefore, that a compromise must be found. Theoretically, it can be shown that low internal resistance is obtainable without entailing excessive inter-electrode capacitances by rendering the current density in the valve high. Hence a highly emissive cathode material such as thoriated tungsten is all to the good. The actual internal resistance of the valves under review is about 200 Ω.

At a filament consumption of about 3.5 kW (17.5 V × 196 A), the saturation current of the cathode is approximately 250 A, but, to allow a safety margin, the maximum values of anode and grid current have so been specified that saturation of the cathode is not by any means reached. The maximum values will be found in *table I*.

**Table I.** Values of the anode current ($I_{a\,\max}$), grid current ($I_{g\,\max}$) and the corresponding anode dissipations ($W_a$) for the transmitting valves TBL 12/100 and TBW 12/100 at which values some quantity reaches its maximum permissible value.

| Application | $I_a$ peak A | $I_g$ peak A | $W_a$ kW | Cooling |
|---|---|---|---|---|
| Telegraphy | 50 | 18 | 36 | air |
| Anode modulation | 72*) | 25*) | 30 | air |
| Class B operation (linear amplifier) | 80*) | 30*) | 95 | water |

*) These values, which are well on the safe side, occur only at the peaks of the deepest possible modulation.

Because of the low internal resistance, losses in the valve are also quite low. In a telephony transmitter the valves will operate at more than 80% efficiency, which is very high for a valve of this output. The value given refers to frequencies up to 30 Mc/s (wavelength 10 m); at higher frequencies the efficiency falls off gradually, although even at 68 Mc/s [7]) it is still some 60% (*fig. 4*).

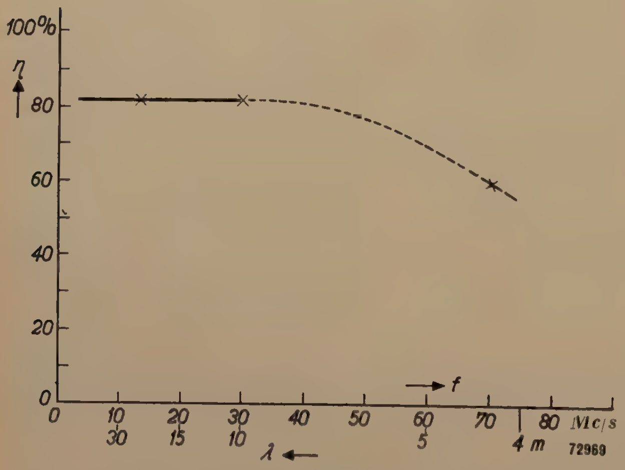


Fig. 4. Efficiency $\eta$ of the TBW 12/100 operating under full load in a telephony transmitter, as a function of the frequency $f$ or the wavelength $\lambda$. The crosses indicate the measured values.

The higher the frequency, the smaller the input power that may be applied; otherwise the displacement currents set up by the valve capacitances tend to overheat the lead-in conductors. When air-cooling is employed, it is a simple matter to arrange for a flow of air to pass through slots in the cooler on to the lead-in conductors of the valve.

Further electrical data relating to these valves are contained in *table II*.

**Table II.** Some electrical data of the transmitting valves TBL 12/100 and TBW 12/100.

| | | |
|---|---|---|
| Filament voltage | | 17.5 V |
| Filament current | | 196 A |
| Amplification factor | | 27 approx. |
| Mutual conductance (max.) | | 120 mA/V approx. |
| Capacitances: | | |
| anode-to-grid | | 86 pF |
| grid-to-cathode | | 116 pF |
| anode-to-cathode | | 3.4 pF |
| Anode dissipation: | | |
| with air-cooling | max. | 45 kW |
| with water-cooling | max. | 100 kW |
| Max. output power | | |
| telegraphy | | 105 kW*) |
| telephony with anode modulation (carrier) | | 65 kW**) |

*) Anode voltage 12 kV
**) Anode voltage 10 kV
For class B amplifiers an anode voltage of 15 kV max. is permissible at frequencies up to 4 Mc/s.

## Cooling

### *Water cooling*

The cooling jacket of a transmitting valve should be capable of conveying away the dissipated energy effectively and economically. In this respect, however, jackets have in the past left a great deal to be desired; with the usual construction water flow was such that only a small amount of turbulence occurred. As a result, only a thin layer of water came into actual contact with the anode surface. In order to keep the temperature of the water below 90 °C (at higher temperatures scale is formed), large quantities of water were needed, viz. about 4 litres/minute per kW dissipated.

A great improvement in economy is effected by arranging for the greatest possible turbulence of the water; each particle of the water then comes in contact with the anode again and again. A description was given in a previous issue of this Review of a jacket in which turbulence is increased by spraying the water against the anode in jets [8]), but in the jacket used with the TBW 12/100 the required result is obtained more effectively still. Owing to the helical ribs *11* on the inside of the jacket

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[7]) In the Philips Laboratories at Eindhoven experiments have been carried out with two TBW 12/100 valves in push-pull serving as the output stage for an experimental television transmitter. Negative modulation was employed, with frequencies up to 68 Mc/s. Output power at the sync. pulse peaks was found to be about 100 kW. The bandwidth of the anode circuit was ample for the 625-line system.
These tests will be described in a later issue of this Review. — Ed.

[8]) M. J. Snijders, A transmitting valve cooler with increased turbulence of the cooling water, Philips tech. Rev. **10**, 239-246, 1949.

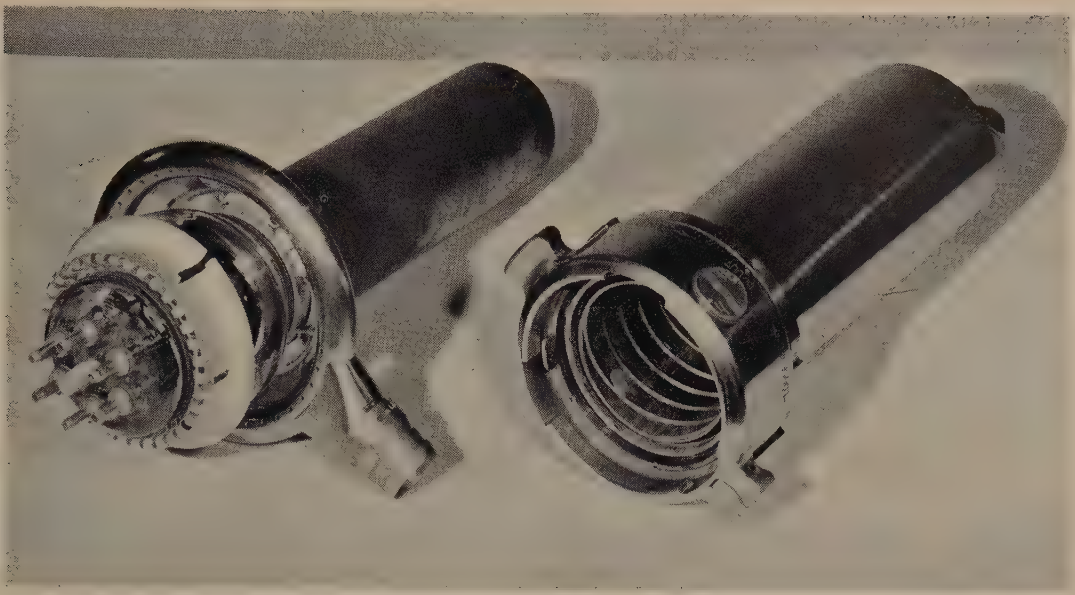

Fig. 5. Left: The TBW 12/100. Right: the water jacket in which the helical ribs — *11* in fig. 3*b* — are clearly visible.

(fig. 3*b* and *fig. 5*), in conjunction with the oblique positioning of the inlet and outlet, the stream of water is primarily helical in shape around the anode, but it also flows between the anode and the ribs, which then function as obstacles, thus ensuring a strong and uniform turbulence. With this jacket less than 1 litre/min per kW is required.

From the practical point of view it is important that the valve can be quickly replaced. With earlier cooling jackets this usually took a considerable amount of time; the maintenance engineer had to loosen a number of bolts with a spanner before the valve could be replaced, and this could not always be done by one man alone. The bolts had then to be tightened again to prevent leakage.

The process is very much simpler in the case of the TBW 12/100; to loosen the valve from the jacket it suffices to turn a ring (*13*, fig. 3*b*) through a small angle by hand; the valve is then tilted out of the jacket and the new one inserted, this being locked in position by turning the ring back. When the valve is inserted, the flange of the anode lies against a packing ring of O-shaped cross-section, and the pressure of the water in the jacket deforms this ring in such a way as to ensure a perfectly tight joint.

For further details of this system, known as the "Grip-o-matic" system, which necessitates no tools whatsoever, we would refer to another publication [9])

Because of the shorter length and lighter weight of the valve (see *table III*), replacement of the

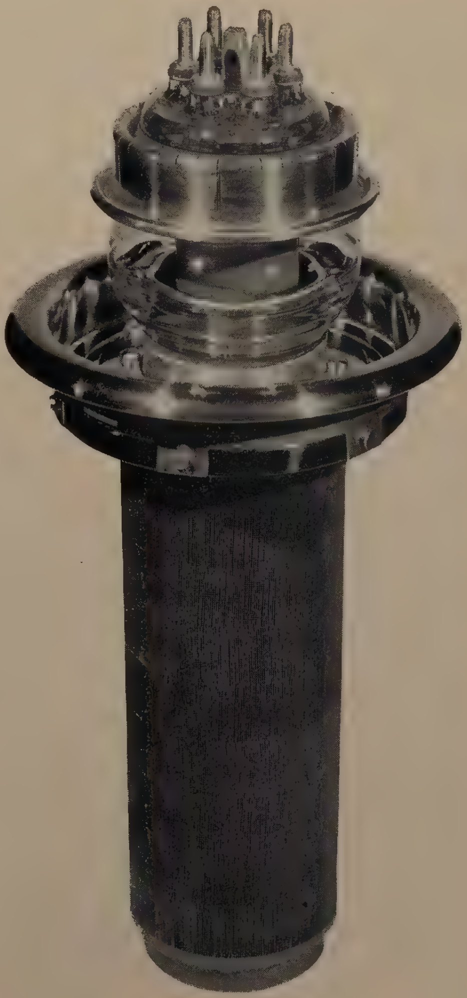

Fig. 6. The TBL 12/100 showing the innumerable fine cooling fins on the anode ("Multifin" system).

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[9]) A. G. Robeer and W. L. Vervest, The "Grip-o-Matic" water jacket for large water-cooled transmitting valves, Communication News **12**, 10-14 1951 (No. 1).

**Table III.** Length and net weight of transmitting valves TA 20/250 and TBW 12/100.

| | Total length (with cooling jacket) cm | Length of anode cm | Weight of valve kg |
|---|---|---|---|
| TA 20/250 | 139 | 70 | 22 |
| TBW 12/100 | 71 | 39 | 14 |

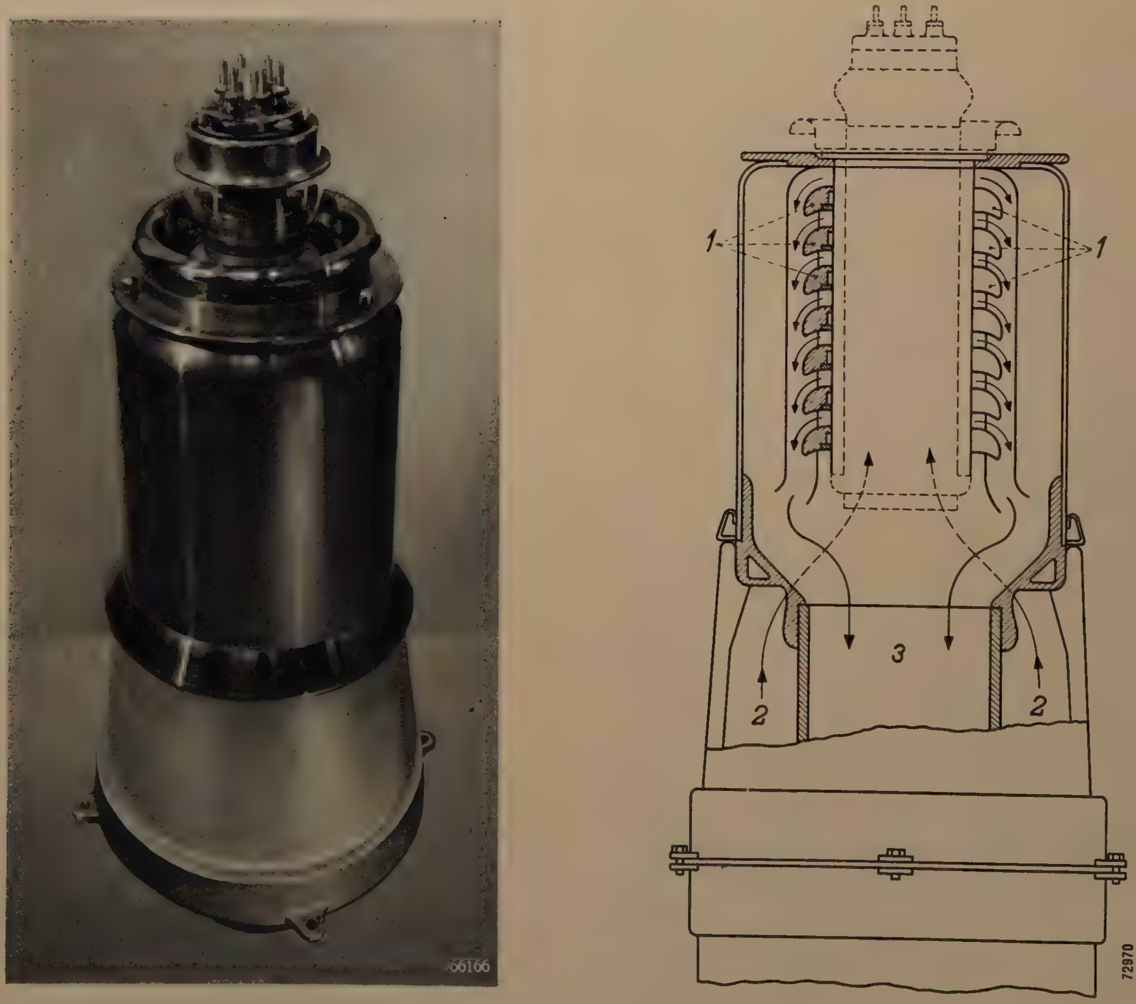
TBW 12/100 is very much easier than that of the TA 20/250 and the chances of accidents during replacements are so much the smaller.

The lower anti-corona ring is made in the form of a tube (see *4*, *5* and *6*, fig. 3*b*) and, at frequencies beyond 6 Mc/s, air is blown into this ring to provide extra cooling; the air streams through a number of holes in the ring and cools the glass-to-metal seals.

### *Air cooling*

The cooler house of the TBL 12/100 is designed on the following principle [10]). It incorporates a very large number of longitudinal cooling fins on the anode surface (*fig. 6*), whence the name "Multifin" system. These fins are very thin and only 1 cm high; they are manufactured on a specially designed folding machine and are made from thin sheet copper, being subsequently silver-soldered to the anode. The resultant cooling area is much greater than that of the heavier fins used in other cooling systems.

A more fundamental difference, however, is to be found in the fact that the air streams in a different direction from that followed in other systems, in which one flow is used, streaming upwards along the cooling fins. Particularly when the anodes are long, the top part usually receives "cooling" air which has already attained quite a high tempera-



*a* *b*

Fig. 7. *a*) The TBL 12/100 in its cooler. *b*) The same in cross-section. The baffles *1* divert the incoming air *2* into a number of short paths which come together in the outlet tube (3).

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[10]) For further details see H. de Brey and H. Rinia, An improved method for the air-cooling of transmitting valves, Philips tech. Rev. **9**, 171-178, 1947.

ture in the lower portions, and the cooling of the upper region is thus not very effective. In the "Multifin" system the incoming air is split up into a number of parallel paths (*fig.* 7), each of which covers only a small zone of the anode, so that each section receives fresh, not pre-heated air.

Here again, care has been taken to secure considerable turbulence; each jet of air is directed radially towards the anode, is then deflected upwards and downwards, and is finally led away radially; the two changes of direction ensure ample turbulence.

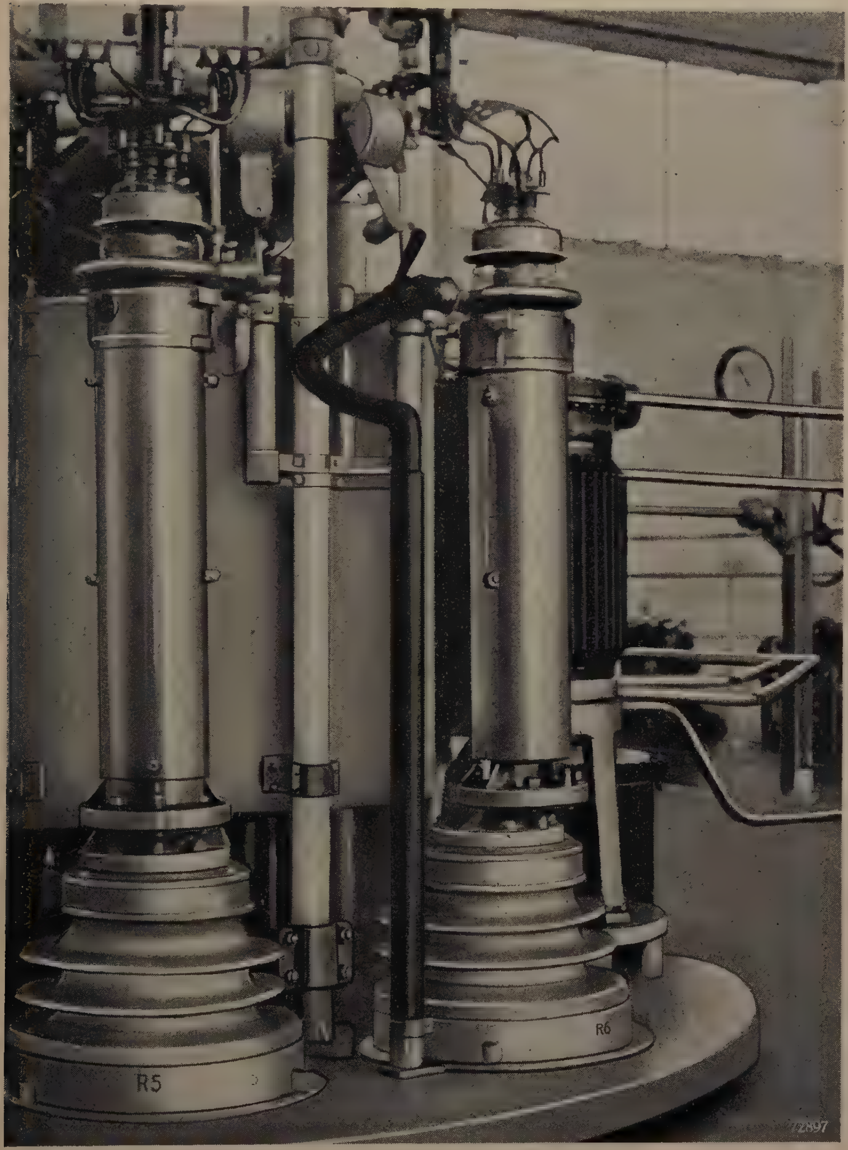

Fig. 8. In the foreground (right) a TBW 12/100 in the 100 kW broadcast transmitter at Mühlacker near Stuttgart.

By this method 50 W to 60 W per cm$^2$ of anode area can be dissipated for a consumption of about 1 m$^3$/min of air per kW, without the temperature of the inner anode wall rising above 180 °C. Compared with the earlier system, this is an improvement in the air consumption by a factor of 2, at the same anode temperature. The used air is led out through a central tube (*3*, fig. 7*b*). This has two advantages in comparison with the older system of merely allowing the air to escape; in the first place the atmosphere in which the valve is mounted is not heated unnecessarily and, secondly, the annoying hiss that always accompanies escaping air is avoided. In broadcasting stations where quiet is essential, the avoidance of unnecessary noise is of course important.

The temperature of the used air is roughly 150 °C and this is a temperature at which the air can often be put in to good use, e.g. for heating the building or, if the valve is employed in a generator for capacitive heating [11]), for pre-heating materials such as plywood etc.

Owing to the presence of the cooling fins, the weight of the TBL 12/100 is 30 kg net, which, although more than that of the TBW 12/100, is not too much for one person to handle.

## Life

Theoretically, a valve of the type TBL or TBW 12/100 would reach the end of its life from one of the following causes:

1) Evaporation of the cathode, more especially of the thorium in the filament.
2) Alloying of the zirconium with the core material of the grid.
3) Saturation of the getters by liberated gases.
4) Deformation of components as a result of recrystallization, sagging of materials carrying relatively heavy mechanical loads, and so on.

Let us now examine these points one by one.

1) With cathodes of pure tungsten operating at a temperature of 2600 °K, evaporation of the tungsten limits the life to some thousands of hours. Thoriated tungsten cathodes work at 2000 °K, and calculations show that it takes about 100 000 hours for all the thorium to evaporate from the filament.

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[11]) See, for example, Philips tech. Rev. **11**, 232-240, 1950.

2) Practical results have shown that after 10 000 working hours there is no sign of alloying of the zirconium on the grid with the core material, at any rate as long as the valve is operated within the limits specified. Tests, however, have shown that overloading can be very damaging in this respect, so that care must be taken not to exceed the maximum ratings.
3) The getters (Zr) are present in such ample quantities that there is no likelihood of their becoming saturated unless the valve develops a leak.
4) It has already been pointed out that in order to minimize the risk of physical distortion, all mechanically loaded components of the valve are made of metals which are not easily deformed, i.e. tungsten and molybdenum. Welded joints, which in the case of these metals are usually very brittle, have been avoided in the design wherever possible; for example all the joints in the grid are effected by riveting or by what is known as the "cut-squash" method.

Ultimately practical tests will have to prove whether the good results to be antipicated from these considerations will be achieved. So far, the results have been very satisfactory, both in broadcasting transmitters and generators, and mention may be made of the Mühlacker broadcast transmitter near Stuttgart (*fig. 8*), where a number of TBW 12/100 valves have been in use under full load since April 1950. The life so far is 12 000 hours and there have been no breakdowns. *Fig. 9* further shows a TBL 12/100 which has been working at the Westerglen (Scottish) transmitter of the B.B.C. since 1st November 1950, also without interruption.

## Economy through higher efficiency

The high output efficiency of the TBL and TBW 12/100 — practically 80% as compared with 65 to 75% in older types — shows a saving of roughly 15 kW per valve on full load, a similar amount being saved by reason of the lower heater power required. On 5000 working hours per year, therefore, the amount of power saved is not less than 150 000 kWh per valve.

At the Westerglen station (50 kW) the introduction of the TBL 12/100 has increased the overall efficiency of the transmitter from 50% to 56%.

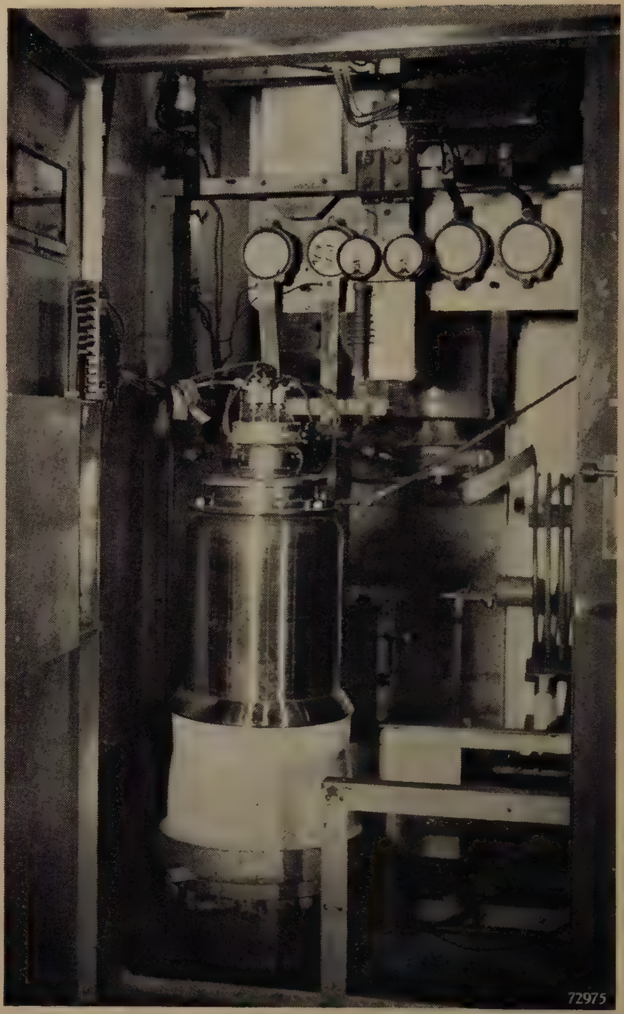


Fig. 9. A TBL 12/100 in the 50 kW Westerglen (Scottish) transmitter of the B.B.C.

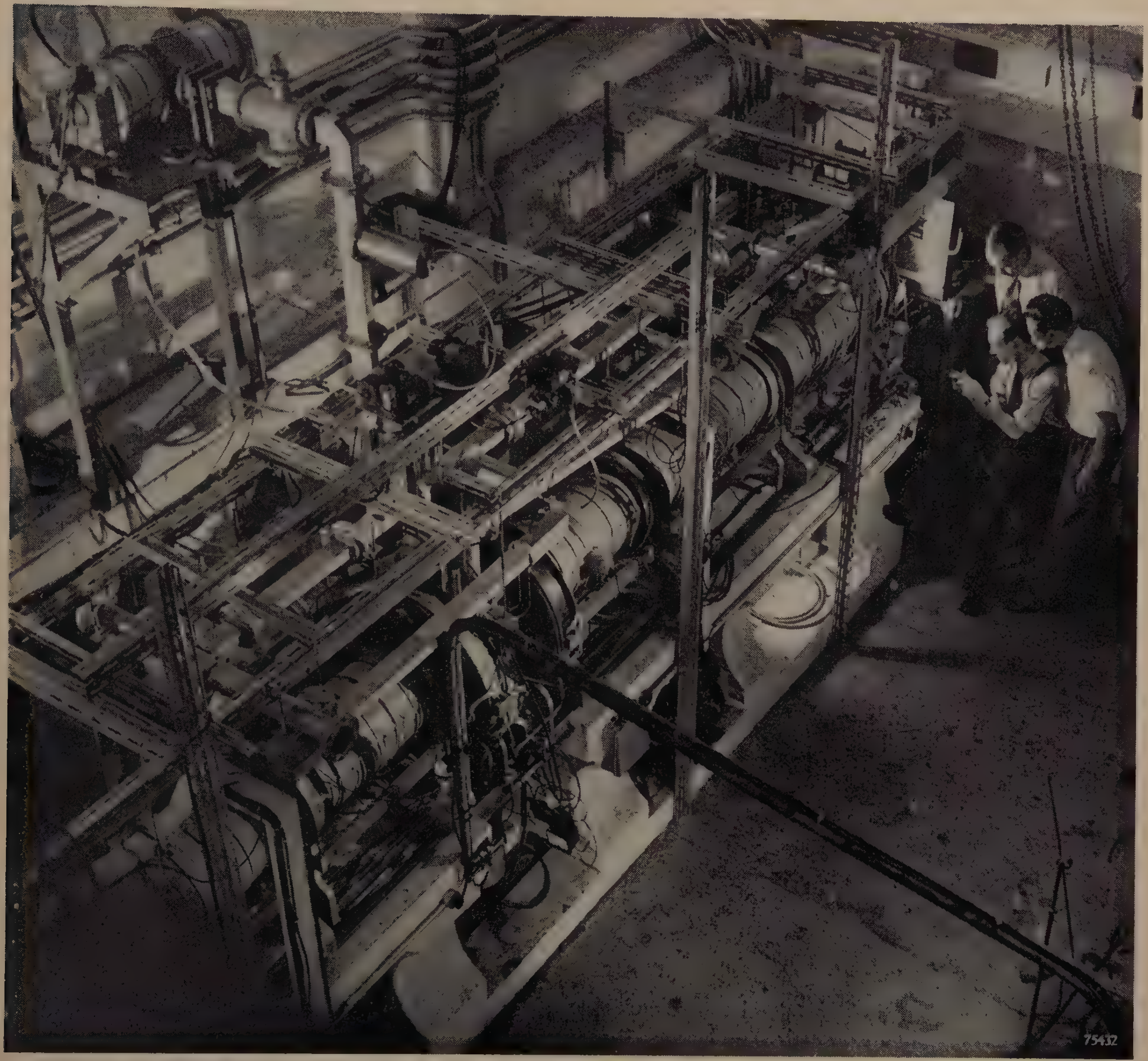
**Summary.** The paper describes two models of a new transmitting triode, the TBW 12/100 and TBL 12/100, the former with water cooling, the latter with forced air cooling. The filament is of thoriated tungsten, which has much better emission properties than pure tungsten, but could not be employed in high-power valves until more efficient getters had been developed. The new valves contain three getters, all of specially prepared zirconium and of which one functions slowly but has a high gettering capacity, one works rapidly, in the form of a layer of zirconium on the grid, and one is specifically for absorbing hydrogen.

The advantages as against comparable valves with pure tungsten filament are: no flashover (Rocky Point effect), thanks to the lower anode voltage required; lower filament consumption (3.5 kW); higher efficiency (a good 80% at frequencies up to 30 Mc/s, approx 60% at 68 Mc/s); longer life, easier replacement because of the lighter weight and smaller dimensions. With 80% output efficiency, an economy of 150 000 kWh per valve and per year of 5000 working hours can be achieved.

At frequencies up to 30 Mc/s the output power is 105 kW (telegraphy) or 65 kW (telephony). For water-cooling, the "Grip-o-matic" water jacket has been introduced, ensuring very high turbulence of the water and a water consumption of less than 1 litre/min per kW to be dissipated.

The air-cooled type is equipped with a very large number of cooling fins ("Multifin" system) of which each section is cooled by separate jets of air. With an air consumption of 1 m$^3$/min per kW, 50 to 60 W per cm$^2$ of anode area can be dissipated. The used air (temperature 150 °C) may be utilised for heating or drying purposes.

# 15 MILLION ELECTRONVOLT LINEAR ELECTRON ACCELERATOR DURING CONSTRUCTION



Photograph Walter Nürnberg

One of the forthcoming issues of this Review will be largely devoted to a description of this electron accelerator, designed and built by a team of scientists at the Mullard Research Laboratories, Salfords, England, under the direction of C. F. Bareford and M. G. Kelliher. The installation was ordered by the Atomic Energy Research Establishment at Harwell and was handed over to them during August 1952, after several months of testing at Salfords.

The photograph pictures the accelerator, whose length is more than 6 metres, during trial assembly in an underground shelter built especially for this purpose at M.R.L. The 10 cm waves which impart their energy to the particles injected into the accelerator are produced by a magnetron which is just visible in the upper left hand corner.

# THE NOISE OF ELECTRONIC VALVES AT VERY HIGH FREQUENCIES

## II. THE TRIODE

by G. DIEMER and K. S. KNOL.

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*The noise of a triode at very high frequencies is determined (as is the noise of a diode) by the form of the current pulses produced in the external circuit by the individual electrons passing through the valve. If the triode is to function satisfactorily as part of an amplifier, the noise produced by the valve in the output circuit must be minimized. This can be achieved satisfactorily by a judicious choice of the elements in the external circuit.*

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### Noise of a triode

Noise of a triode is caused, as in the case of a diode, by fluctuations in the state of motion of the electrons as they pass through the valve. As explained in a previous article [1]), the noise phenomenon can be investigated by resolving the convection noise current in the valve into its components at various frequencies (Fourier analysis), and investigating the relationship existing between these convection current components and the noise current components in the external circuit of the valve. The convection current is defined as the net amount of charge passing a given cross-section of the valve per second. At frequencies of the order of the reciprocal of the transit time of the electrons, the convection current is related only indirectly to the current in the external circuit, since the latter is actually an induced current (the electrons induce charges on the electrodes on their way through the valve). The current in the external circuit, or at least its Fourier components, can, however, be expressed in terms of the Fourier components of the convection current which starts at the cathode, by means of the transit time function $\Psi$. This function depends on the transit time $\tau$ of the electrons between the planes of two electrodes in the valve, and, of course, on the frequency $f$.

Similarly, if a signal voltage is present in the input circuit of the valve, it is possible to express the signal voltage in the output circuit in terms of the convection signal current produced by the signal, by means of a transit time function. This convection current passes, however, in an electric field which has a certain phase relation with the externally applied signal. There is obviously no such relationship for noise currents. As a consequence, the transit time function for the signal currents differs from that for the noise currents.

In a triode two "interelectrode spaces" may be distinguished, namely the cathode-to-grid space and the grid-to-anode space (cg-space and ga-space respectively). In order to investigate the noise and signal currents, the convection currents in both spaces must be considered in turn. It will be seen that a relationship exists between the two currents, a relationship depending on the nature of the circuit in which the triode is employed.

### Triode circuits

Three different types of circuit are possible for a triode, in which each electrode in turn is the common electrode of the input and output circuits. These three circuits are known as: grounded-cathode, grounded-grid and grounded-anode circuits. They are shown diagrammatically in *fig. 1*; *1* and *2* indicate input and output circuits respectively, which for the moment will be considered to be short-circuited for high frequencies, so that H.F. currents may flow in these leads but no H.F. voltages occur between the planes of the electrodes.

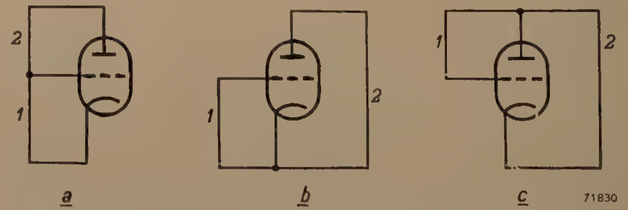

Fig. 1. Schematic representation of the three ways of connecting a triode: (*a*) grounded-grid circuit; (*b*) grounded-cathode circuit *c)* grounded-anode circuit; *1* is the input circuit, *2* the output circuit, both considered to be short-circuited for H.F. currents.

For low and medium frequencies, the grounded-cathode circuit is usually employed, whilst the grounded-grid circuit is preferred in amplifiers for very high frequencies. In the latter circuit the grid

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[1]) G. Diemer and K. S. Knol, The noise of electronic valves at very high frequencies. I. The diode, Philips tech. Rev. 14, 153-164, 1952 (No. 6) This article will be referred to as I in the following.

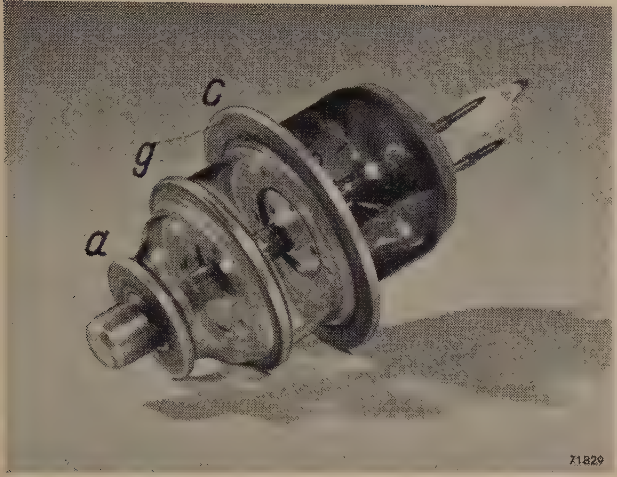

Fig. 2. Experimental disc-seal triode: *c* is the cathode disc, *g* the grid disc and *a* the anode disc (fitted with a cap).

acts (when it is earthed for high frequencies) as a natural screen between the input electrode (cathode) and the output electrode (anode). Moreover, in grounded-grid circuits, if the grid is mounted on a metal disc which extends outside the glass envelope ("disc-seal" triode, *fig. 2*), the H.F. input currents and the H.F. output currents are not able to influence each other, since these currents flow on opposite sides of the grid disc as a consequence of the skin effect. In this way the electric and magnetic couplings, which could cause feed-back, is made as small as possible. The grounded-anode circuit is seldom employed, and in the following it will be left out of consideration.

### Current pulses in a triode

The potential distribution between the cathode and grid planes in a triode differ but little from that in a diode if the grid and anode voltages are not extremely high. There is a potential minimum located somewhere between cathode and grid. For the sake of simplicity, this potential minimum will be considered to be immediately adjacent to the cathode (as, in fact, it frequently is). An electron passing from cathode to grid (transit time $\tau_{cg}$) produces, precisely as in the case of the diode, an almost parabolic current pulse in the external circuit, in this case the lead between cathode and grid; this is the input circuit in both grounded-grid and grounded-cathode circuits. This current pulse is shown graphically as curve *1* in *fig. 3a*. After passing the grid, the electron produces, during its passage from grid to anode (transit time $\tau_{ga}$), a current pulse as shown by curve *2* in fig. 3*a* in the lead between the two electrodes, i.e., in the case under consideration, the output circuit. In the grounded-grid circuit the latter pulse flows only in this output lead. In the grounded-cathode circuit, however, the second pulse flows through the input and ouput leads in series, the current direction in the input circuit being opposed to that of the first pulse in this lead. Thus, whilst fig. 3*a* gives the current pulses in input and output circuits for grounded-grid circuits, the total current pulses in both circuits of the grounded-cathode circuit is as shown in fig. 3*b*.

The noise of a triode will now be derived from these current pulses.

The current pulses in the input lead are responsible for the noise in this lead, the so-called input noise. Similarly, the current pulses in the output circuit are responsible for the output noise. As will be shown later, it is not sufficient to study only the output noise; if the input circuit contains finite impedances, the input noise is also of importance in defining the properties of the valve and circuit. It will therefore be useful to discuss both input and output noise.

The two current pulse curves *1* and *2* in fig. 3 obviously have an area equal to the electronic charge. If a Fourier analysis is made of the current pulses in the input lead of a grounded-cathode circuit, it is found that at low frequencies the two pulses practically neutralise each other, in other words, the amplitudes of the input components at low frequencies are practically zero. This is not the case for grounded-grid circuits; at low frequencies, however, the pulses may be considered to be infinitely small, giving for these frequencies a spectrum with

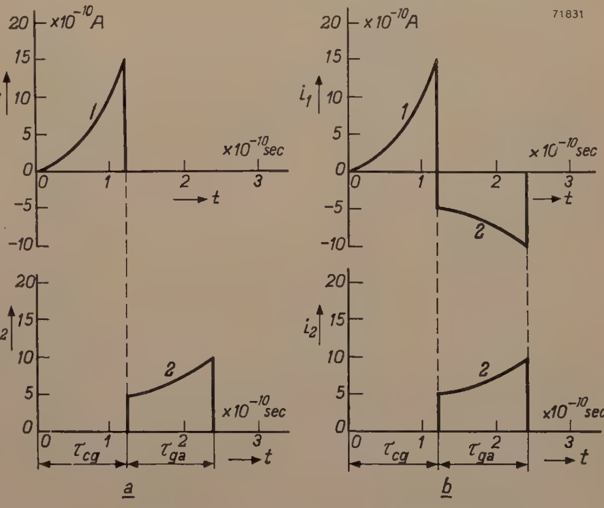

Fig. 3. (*a*) An electron moving from cathode to anode in a triode produces current pulses in the input circuit ($i_1$) and the output circuit ($i_2$). With the simplifying assumption that the potential distribution between cathode and grid is such that Child's law applies (see I), these pulses, in the grounded-grid circuit, have the shape *1* (in the input circuit) and *2* (in the output circuit). (*b*) In the grounded-cathode circuit, in addition to the pulse *1*, a pulse *2* flows through the input circuit (but in the opposite direction). Only pulse *2* flows in the output circuit.

a finite amplitude, independent of frequency (see I). This is true also for the output currents in both circuits. In other words, at low frequencies the output noise in grounded-cathode circuits, and both input and output noise in grounded-grid circuits, are directly comparable with the normal shot effect in diodes. The input noise in the grounded-cathode circuit is practically zero.

We proceed to investigate noise at frequencies of the order of the reciprocal transit time (or less). If the input and output circuits are still considered as being short-circuited, there exist certain phase relationships between the noise currents in the input and output leads. First, the currents in the input lead will be investigated by means of a vector diagram (*fig. 4*). The arrows in this diagram represent the components of the various noise current components at a given frequency $f$.

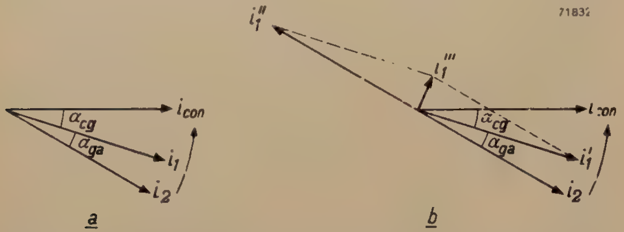

Fig. 4. Vector diagram for the noise currents in a triode. The vectors of the current components at a given frequency are shown, the curved arrow points in the direction of rotation, the angle between two vectors is the phase difference of these current components. $i_{con}$ represents the convection noise current of the electrons which start their trajectory through the valve at the cathode; in drawing the amplitude of $i_{con}$, account has been taken of the noise suppression due to the space charge; $i_1$ is the noise current in the input lead of the triode, $i_2$ that in the output lead. The angle $\alpha_{cg}$ between $i_1$ and $i_{con}$ is proportional to the transit time $\tau_{cg}$ in the cathode-to-grid space, the angle $\alpha_{ga}$ between $i_2$ and $i_1$ is proportional to $\tau_{ga}$. (*a*) applies to grounded-grid circuits; (*b*) to grounded-cathode circuits. In the latter circuit the current in the input circuit consists of two components $i_1'$ and $i_1''$ corresponding to the pulses *1* and *2* in fig. 3*b*; $i_1'$ lags $i_{con}$ by the phase angle $\alpha_{cg}$, $i_1''$ is 180° behind $i_2$ (the current in the output circuit) and has the same amplitude as $i'_1$. The total noise current in the input circuits is $i_1'''$. For low frequencies $\alpha_{ga}$ is almost zero and $i_1'''$ is then also zero. Only at high frequencies has $i_1'''$ a finite value, albeit small, and then leads practically 90° with respect to $i_1'$.

In grounded-grid circuits the phase of the noise current $I_1$ in the input lead lags by a certain angle, the transit time angle $\alpha_{cg}$ (related to $\tau_{cg}$ and the frequency by a transit time function), behind the convection current $i_{con}$ which starts at the cathode (the potential minimum). The noise current $i_2$ in the output lead lags behind $i_1$ by a transit time angle $\alpha_{ga}$. The resultant noise current in the input lead is in this case given directly by $i_1$, the magnitude of which is almost independent of frequency for medium frequencies.

The case of the grounded-cathode circuit is different. The vector diagram corresponding to this circuit is given in fig. 4*b*. Firstly, a noise current $i_1'$ flows in the input lead lagging in phase by an angle $\alpha_{cg}$ behind $i_{con}$, and, secondly, a noise current $i_1'''$ which is in anti-phase with the noise current $i_2$ in the output lead. The latter lags by a further angle $\alpha_{ga}$ behind $i_1'$. The resultant noise current in the input lead is thus $i_1'''$. A comparison of figures 4*a* and 4*b* clearly shows that in grounded-cathode circuits the resultant noise current in the input lead is much smaller than in grounded-grid circuits. At low frequencies the input current in grounded-cathode circuits is, as previously shown, almost zero. (This is because $\alpha_{ga}$ is practically zero at low frequencies.) It is true that at high frequencies a finite input noise current occurs, but this is still quite small. It is then seen that even at high frequencies the pulses *1* and *2* of fig. 3*b* cancel one another to some extent.

The contribution of the shot effect to the noise current at high frequencies in the input lead of a triode is called the induced grid noise (or the transit time noise) of the triode [2]). It has been shown that this noise is much smaller for grounded-cathode circuits than for grounded-grid circuits. These induced grid currents have no influence on the output noise currents, so long as the input is short-circuited for H.F. currents. The influence at finite values of the input admittance is dealt with below. It will be seen that this influence is not always harmful.

If an H.F. signal voltage $V_g$ is applied to the grid of a triode, a further vector diagram may be constructed for the Fourier components of the signal currents corresponding to the frequency of the voltage applied, now however with transit time angles: $\alpha'_{cg} \neq \alpha_{cg}$ and $\alpha'_{ga} \neq \alpha_{ga}$, since the transit time functions are different (*fig. 5*). As can be seen, the component $I_1'''$ of the input current which is in phase with the input signal is much smaller in the case of grounded-cathode circuits than in the case of grounded-grid circuits. This means that the electronic input damping, at least for medium frequencies, is smaller in the former case. At frequencies of the order of the reciprocal transit time, the damping in the case of grounded-cathode circuits increases. This is known as transit time damping.

This transit time damping may be regarded as an fictitious noise source producing the transit time noise (i.e. the high-frequency noise in the input circuit). In this case the equivalent temperature of the "transit time resistance" appears to be about

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[2]) This is an extension of the similarly named concept introduced by C. J. Bakker: Philips tech. Rev. 6, 129-138, 1941.

1.5 times the cathode temperature. It is, however, unsound to introduce the concept of such a transit time resistance, since the phase relations between the noise currents in the input and output circuits are then neglected. Such a correlation between input and output currents would apply fully only if all the electrons had the same transit time. Due to the irregularities in the cathode surface, the velocity distribution of the electrons, and the inhomogeneity in the potential distribution introduced by the grid wires, there appears, however, a spread in the transit time. The correlation is then no longer complete, but does not wholly disappear [3]).

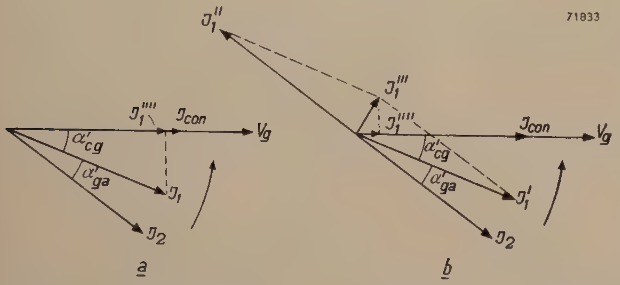

Fig. 5. Vector diagrams for the signal currents in a triode. The symbols have the same meanings as in fig. 4 (to show that signal currents are now referred to, the small letters are replaced by capitals and the transit time angles are dashed). The convection current which starts at the cathode, $i_{\text{con}}$, is in phase with the signal voltage $V_g$. (*a*) Grounded-grid circuit. In the latter circuit the component $I_1''''$, which is in phase with the signal voltage (and thus with $I_{\text{con}}$) is very small: there is only slight electronic input damping.

For the same reason, it is undersirable to make use at high frequencies of the concept of equivalent noise resistance, i.e. a resistance that is imagined to exist in the input lead of a hypothetical noise-free valve, and that is just high enough to produce at room temperature the same noise currents in the output lead as are in fact produced by the valve itself. It can be shown that this fictitious noise resistance is closely related to the mutual conductance of the valve.

The mutual conductance is determined by those electrons that leave the cathode with forward velocities falling within a narrow velocity range of such values that their behaviour (passing the potential minimum or not, and thus contributing to the anode current or not) can be entirely determined by the variations in depth of the potential minimum, such as may be caused by grid voltage variations. In Part I these electrons were termed $\alpha$-$\beta$ electrons (they are also responsible for the static internal resistance of a diode).

Looked at in this way, it would appear that the noise in the output circuit is determined by the $\alpha$-$\beta$-electrons. From the above, however, it will be clear that this noise is caused almost wholly by the much larger number of electrons which always reach the anode (called $\beta$-electrons in Part I). If only these $\beta$-electrons are considered together with their influence on the potential minimum, the output noise can be computed without the mutual conductance being taken into consideration. The output noise, therefore, depends only in a very indirect way on the mutual conductance, and on the equivalent noise resistance. It is therefore apparent that, if the noise is defined by means of such a resistance, the relationship between the various causes of the noise is very easily overloocked.

## The triode in amplifier circuits; noise factor

Attention will now be paid to the influence that the electronic noise of a triode has on the useful gain that can be given by the valve, when used in an amplifier stage, the valve circuits thus not being short-circuited. First, the concept of useful gain will be considered. A measure of the relative influence of the noise is given by the signal-to-noise ratio, by which is meant the ratio of the signal power to the noise power, measured in the direction from the input to the output through a given "cross-section" of the amplifier (e.g. the output terminals).

The possibility of obtaining effective gain of a given small signal is entirely dependent on the signal-to-noise ratio. If the signal is so weak that the signal-to-noise ratio at the output terminals is smaller than unity, the output signal is "drowned" by the noise and becomes very difficult or impossible to separate from the noise. In practice, the tolerable signal-to-noise ratio at the output side of the amplifier depends on the nature of the application. For television receivers, for example, a ratio of the order of 100 is desirable, whilst for radar apparatus a signal-to-noise ratio of just over 1 is considered satisfactory. In special amplifiers, as used, for example, for measuring noise from inter-stellar space at decimetric waves, one must be satisfied with signal-to-noise ratios of the order of 1%. It has been found, however, that by special measuring procedures, variations in the average value of the "signal" to be determined (which in this case has itself a noise character) of approximately 0.1% of the amplifier noise can still be detected.

At very high frequencies, the signal-to-noise ratio at the output of the amplifier is affected by the circuit elements external to the valve [4]). This is not the case at low frequencies. The cause can be found in the different behaviour of the signal currents and noise currents on their way through the valve, as previously mentioned. If there is an admittance (such as is caused by the presence of feedback) in the circuit between input and output terminals, any noise voltage existing at the input terminals can cause a corresponding noise voltage at the output terminals via two different paths, i.e. through the valve itself and by-passing the valve. A signal in the input similarly makes two contributions to the signal in the output circuit. But, at high frequencies, the signal currents and noise currents are, on their way through the valve, affected in different ways (different transit time functions). Thus, variation of the feedback admittance will change the signal-to-noise ratio at the output terminals.

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[3]) A. van der Ziel and A. Versnel, Philips Res. Rep. **3**, 255-270, 1948; W. Kleen, Frequenz **3**, 209-216, 1949; A. van der Ziel, Can J. Technology **29**, 540-553, 1951.

This point will be reverted to presently, but it might be remarked here that there is a possibility of minimising the noise in the output lead, and thus making the signal-to-noise ratio at the output terminals very large indeed. This may be achieved, for example, by so choosing the feedback admittance that advantage may be taken of the phase relations of the various noise currents. This possibility (which is almost always denied by current opinion) exists for both grounded-grid and grounded-cathode circuits.

The influence of the feedback admittance on the noise in the output circuit due to the phase relations existing between the various noise currents, emphasizes once again the undesirability of describing the H.F. output noise by the concepts of equivalent noise resistance and noise temperature.

The signal-to-noise ratio at the output terminals of the amplifier depends on the input signal. The latter may be varied over a wide range, and thus also the signal-to-noise ratio, without necessarily any change in the noise. In order to determine quantitatively the sensitivity of an amplifier (defined in this case as the weakest signal that may still be "usefully" amplified) the signal-to-noise ratio at a standard level of the input signal power should be known. The signal power which is equal in strength to the maximum noise power that can be produced by an arbitrary resistance in thermal equilibrium with the surrounding (i.e. $kT\Delta f$, see equation (2) in I) may, for example, serve this purpose; such a resistance is always present in the input circuit. The signal-to-noise ratio corresponding to this standard input signal is known as the noise factor $F$ [5]). The noise factor is frequently expressed in decibels, and is then known as the noise figure $N$.

In the literature definitions of $F$ are often seen which, at first sight, seem to differ one from the other. In order to conform as closely as possible with modern views, the above definition will be generalised (without, however, becoming at variance with it), as follows: $F$ is the quotient of the signal-to-noise ratio at the terminals of the input resistance at room temperature — the latter being considered to produce pure thermal noise — and the signal-to-noise ratio at the output terminals of the amplifier. The previous definition is a special case of this definition: i.e. the standard signal was chosen so that the signal-to-noise ratio at the terminals of the input resistance was equal to unity.

Once the noise factor of an amplifier, defined in this way, is known, it is possible to compute directly, without further reference, how small the input signal to be amplified may be, if the signal-to-noise ratio is not to drop below the limit specified for the case under consideration. At not too high frequencies, the noise figure of a good radio or television amplifier must not exceed a few decibels.

As long as the amplifier amplifies linearly, $F$ is independent of the amplitude of the signal. In the determination of $F$, the signal is frequently increased to such an extent that the signal-to-noise ratio at the output of the amplifier is equal to unity. In this case $F$ is equal to the ratio of the noise power fed to the input to the noise power of the (thermal noise producing) input resistance. If all the noise is imagined to be produced in an input resistance which is supposed to have a given, fictitious temperature, this temperature must be equal to $F$ times the room temperature. For an ideal amplifier, $F$ should be equal to unity.

At high frequencies great difficulties are encountered in the measurement of $F$. For example, even in a single stage amplifier, the noise produced by the measuring instrument used to determine the signal-to-noise ratio at the output terminals influences the determination of $F$. Again, the measured thermal noise of a resistance depends on the bandwidth of the amplifier. Further, in the case of superheterodyne amplifiers (in agreement with the definition) when considering the bandwidth, it is necessary to take into account also the image frequencies in the frequency characteristic as passed by the input of the amplifier at the input side. If this is done, the noise factor is then roughly twice as great as when the image frequencies are left out of consideration (which is in better agreement with the sensitivity found in practice).

It can easily be shown, if the signal source is not matched to the input of the amplifier (that is, if the internal resistance of the signal source differs from the input resistance of the amplifier), and the power supplied by the signal source is thus not at its maximum, that the correct value of $F$ can be calculated by using in the definition that value of the input signal-to-noise ratio which would apply if the matching were correct.

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[4]) K. S. Knol and A. Versnel, Physica **15**, 462-464, 1949.

[5]) H. F. Friis, Proc. Inst. Radio Engrs. **32**, 418-423 and 729, 1944.

The noise level reached in the output circuit of an amplifier when finite admittances are present in the valve circuit will now be investigated, and at the same time the possibility, already mentioned, of achieving a low noise factor will be considered in greater detail. Only a few examples will be dealt with.

### Amplifier with grounded-grid circuit, without feedback admittance

*Fig. 6* shows an amplifier stage with a grounded-grid circuit. The valve admittances are separated into electronic admittances (with suffix *e*, indicated by broken lines) and non-electronic admittances, due, for example, to resistance losses in the leads. For the present, non-electronic feedback admittance will be assumed to be absent. First the dependence of the output signal on the input matching will be investigated. The signal source (aerial) is inductively coupled (ratio of transformation $t$) to the input circuit, and supplies to it an electromotive force $tV$. The input admittance due to the coupling with the signal source is equal to $g/t^2$, in which $g$ is the conductance of the source; this is in series with the impedance $Y_1$. The admittance between grid and anode is denoted by $Y_2$. It can be shown that the available signal power at the output increases with $t$ (in spite of the fact that the input voltage then decreases), since in grounded-grid circuits the output admittance depends largely on $t$[6]). The output noise current may be studied as follows, by means of a vector diagram (*fig. 7*). Two types of convection currents originate from the cathode (potential minimum), i.e. the primary noise currents, described by $2eI_a\Delta f\Gamma^2$ (see I, formula (6) where $\Gamma$ is the noise-suppression factor of the space charge) and secondary noise currents, resulting from induced fluctuations $\Delta V_g$ in the grid voltage. These latter are caused by the noise currents flowing in the external input admittance: $\Delta V_g = -i_1/(Y_1 + g/t^2)$. The total convection noise current originating at the cathode is represented by the vector $i_{con}$. The noise current $i_1$ in the external input circuit lags behind $i_{con}$ by the transit time angle $\alpha_{cg}$, and $\Delta V_g$ is in its turn in anti-phase with $i_1$ if $Y_1$ is taken to be real and $g/t^2$ is small (i.e. for large values of $t$; in this case the most favourable noise suppression is obtained). The output noise current is made up of $i_2$ (lagging behind $i_1$ by the angle $\alpha_{ga}$) and the secondary noise current $i_2'$ (caused by $\Delta V_g$), which lags behind $\Delta V_g$ by the angle $\alpha_{cg}' + \alpha_{ga}'$. Herein $\alpha_{cg}'$ and $\alpha_{ga}'$ are "signal" transit angles, since $i_2'$ is caused by $\Delta V_g$, so that it is affected in the same way as a signal current. At low frequencies, i.e. at medium large transit time angles, the phase difference between $i_2$ and $i_2'$ can be made practically equal to 180° ($\alpha_{cg}'$ and $\alpha_{ga}'$ then do not differ greatly from $\alpha_{cg}$ and $\alpha_{ga}$ respectively), so that the resultant output noise current $i_2'' = i_2 + i_2'$ can become almost zero at high ratios of transformation [7]). The noise factor is then at its minimum value. This is illustrated in *fig. 8*. The measurement of $F$ as a function of the ratio of transformation $t$ is not carried out in practice by placing a transformer between the aerial and the input circuit, but by shunting a resistor $R_a$ equivalent to the reflected aerial resistance, i.e. $t^2/g$, directly across the input circuit. In fig. 8, $F$ is plotted against the value of this resistor for the case of an experimental shortwave triode. The measurements were carried out at a wavelength of 7 m. The very low minimum value of $F$ is reached when the reflected aerial resistance

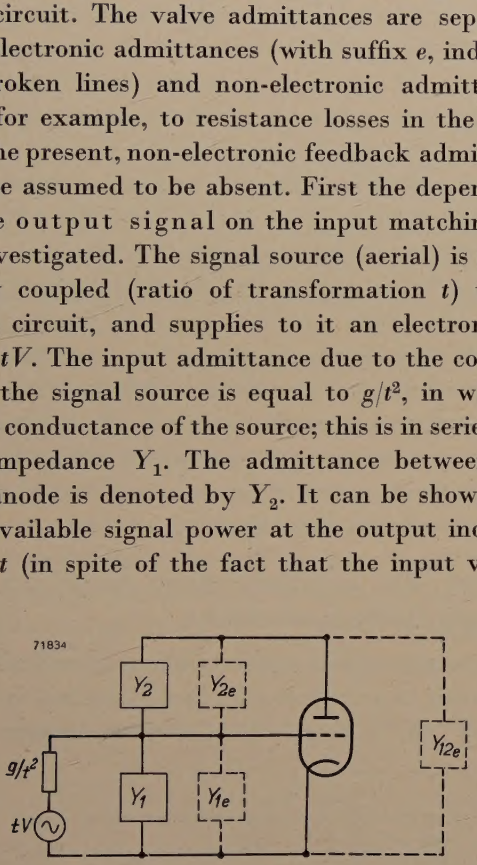

Fig. 6. Basic circuit of an amplifier stage with grounded-grid circuit without feedback of a non-electronic nature. The various admittances are split into electronic (drawn in broken lines and with index $e$) and non-electronic admittances; $Y_1$ and $Y_{1e}$ are the input admittances, $Y_2$ and $Y_{2e}$ output admittances, $Y_{12e}$ electronic feedback admittance.

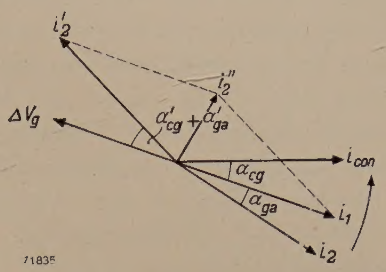

Fig. 7. Vector diagram of the noise currents in the output circuit of a triode in a grounded-grid circuit. The fluctuations of the grid voltage $\Delta V_g$, which are in anti-phase with the input noise current $i_1$, produce a secondary output noise current $i_2'$, which at not too high frequencies can partially compensate the primary output noise current $i_2$ (resultant current $i_2''$). $i_1$ is the input noise current, $i_{con}$ the convection current which starts at the cathode.

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[6]) A. van der Ziel, Philips Res. Rep. **1**, 381-399, 1946.

[7]) A. van der Ziel and A. Versnel, Philips Res. Rep. **3**, 255-270, 1948; A van der Ziel and K. S. Knol, Philips Res. Rep. **4**, 168-179, 1949.

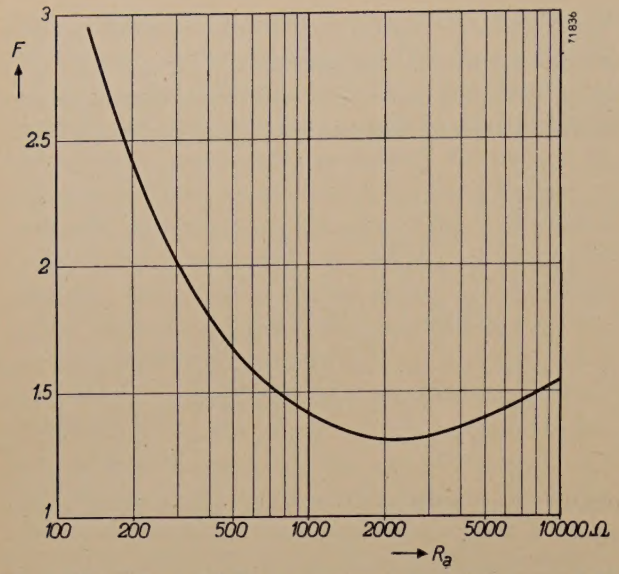

Fig. 8. Noise factor $F$ of an experimental triode in grounded-grid circuit as a function of the reflected aerial resistance $R_a$, measured at a wavelength of 7 m. $F$ is at its minimum when $R_a$ is about 2000 Ω. With an aerial resistance of 70 Ω this corresponds to a ratio of transformation of 5.4.

is about 2000 Ω. At an aerial resistance of 70 Ω this corresponds to a ratio of transformation of $t$ = 5.4.

At centimetric wavelengths a number of effects appear which cause $F$ to increase. These effects are as follows:

1) The total-emission noise (i.e. the noise originating in the input circuit due to the electrons which recoil before the potential minimum, the so-called $\alpha$-electrons, see I) commences to make an important contribution to the induced input noise.
2) The total-emission damping, i.e. the damping caused by the $\alpha$-electrons, reduces the gain.
3) The phase relations between the various noise currents become such that hardly any compensation exists (very large transit angles).
4) The clearance between grid and cathode in valves used for centimetric waves is frequently so small that the potential minimum lies close to the grid, so that the controlling action of the grid wires decreases. The spread in transit times becomes much larger, since the distances between the grid wires are no longer small in comparison with the clearance between grid and cathode.
5) The feedback between the output and input circuits increases, in general, strongly [8].
6) The valve conductances increase, due to increases in skin-effect losses and dielectric losses. This results in a reduction of the gain [9].

By using a special technique in the manufacture of triodes, it is possible to achieve very small inter-electrode clearances and high current densities. The effects mentioned above are then largely overcome. In this way noise figures in the centimetric wave range can be obtained which are as low as those for normal broadcast valves for metric waves. Philips triode EC 80, of fairly normal construction and pin leads [10], has a noise figure of 7 dB at $\lambda = 2$ m. With an experimental disc-seal triode such as that shown in fig. 2, it is possible to reach at $\lambda$ = 10 cm a noise figure of about 10 dB with a power gain of about 10 dB [11]. At $\lambda$ = 75 cm, it was shown that the noise figure in a grounded-grid circuit is larger than that in a grounded-cathode circuit [12]. This result indicates that for extremely low noise figures the grounded-cathode circuit is to be preferred, provided the gain can remain at a satisfactorily high value and there is no risk of the amplifier becoming unstable.

These very low noise figures can be reached by taking advantage of the correlations that exist between the various noise currents. The influence of a feedback admittance in an amplifier with grounded-cathode circuit will now be discussed, these correlations being taken into account.

## Amplifier with grounded-cathode circuit and non-electronic feedback admittance

The output of an amplifier with a grounded-cathode circuit is assumed to be short-circuited for H.F. currents (*fig. 9*). By means of a vector diagram once

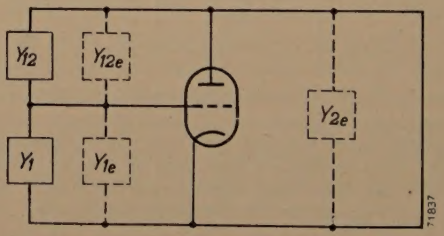

Fig. 9. Basic circuit of an amplifier stage with grounded-cathode circuit and finite feedback admittance, with short-circuited output. The symbols have the same meanings as in fig. 6.

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[8] S. D. Robertson, Bell Syst. Techn. J. 28, 647-655, 1949; G. Diemer, Philips Res. Rep. 5, 423-434, 1950.

[9] K. Rodenhuis, Philips Res. Rep. 5, 46-77, 1950; G. Diemer and K. Rodenhuis, Philips Res. Rep. 7, 36-44, 1952 (No. 1).

[10] A description of this valve may be found in: K. Rodenhuis Two triodes for reception of decimetric waves, Philips tech. Rev. 11, 79-89, 1949.

[11] In a provisional report (Philips Res. Rep. 5, 153-154, 1950) 7 dB was given as the lowest value of $N$. These measurements proved later to contain some inaccuracies.

[12] These measurements were carried out by Dr. Stumpers of this laboratory.

again, the behaviour of the noise currents (*fig. 10*) will be studied. Consider a small convection current $i_{\mathrm{con}}$ originating at the cathode; this current represents the resultant of all the convection currents contributing to the noise. The current primarily excited in the output circuit is $i_2$; because of the input admittance $Y_1$, in which for convenience the reflected aerial resistance $R_{\mathrm{a}} = t^2/g$ is included, there is a small resultant noise current $i_1'''$ which leads $i_2$ in phase by approximately 90° (see fig. 4*b*). This is quite different from the case of the grounded-grid circuit, in which the resultant noise current in the input circuit leads $i_2$ by the small angle $\alpha_{\mathrm{ga}}$ only (fig. 7). The noise current $i_1'''$ produces a grid voltage variation $\Delta V_{\mathrm{g}} = -i_1'''/Y_1$; the phase of $\Delta V_{\mathrm{g}}$ is controlled with respect to $i_1'''$ by the choice of $Y_1$. The most favourable noise suppression is obtained when $Y_1$ is purely capacitive (thus $t$ must again be large). The phase difference between $\Delta V_{\mathrm{g}}$ and $i_1'''$ is then 90°, and the difference between $\Delta V_{\mathrm{g}}$ and $i_2$ is almost 180°. This is an analogous situation to that reached with a grounded-grid circuit when a real value was chosen for $Y_2$. The grid voltage variation produces a further current in the output circuit lagging almost as much in phase behind $\Delta V_{\mathrm{g}}$ as $i_2$ lags behind $i_{\mathrm{con}}$, i.e. the angles $(\alpha_{\mathrm{cg}}' + \alpha_{\mathrm{ga}}')$ and $(\alpha_{\mathrm{cg}} + \alpha_{\mathrm{ga}})$ respectively. For medium transit angles, just as with a grounded-grid circuit, it is possible also with a grounded-cathode circuit, by suitable choice of $Y_1$, to make the angle between $i_2$ and $i_2'$ almost equal to 180° and to give the two currents almost the same amplitude, so that the resultant output noise current $i_2''$ is very small. With the compensation thus achieved the gain is also decreased (the matching of the input circuit to the aerial is less favourable since $t$ had to be $\gg 1$), but since the noise decreases relatively more quickly, the noise factor decreases. This is illustrated in *fig. 11a*, which relates to an EF 50 pentode used as a triode at a wavelengh of 1 m. The feedback admittance which is a consequence of the capacitance between grid and anode is neutralised by shunting this capacitance by an inductance. The feedback admittance is thus reduced almost to zero. Curve *1* shows the noise factor $F$ as a function of

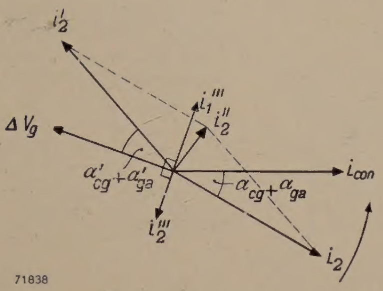

Fig. 10. Vector diagram of the noise currents for a triode in a grounded-cathode circuit. $i_{\mathrm{con}}$ is the noise current which starts at the cathode. This results in a primary noise current $i_2$ in the ouput lead. The resultant noise current $i_1'''$ in the input lead produces grid voltage fluctuations $\Delta V_{\mathrm{g}}$, which, in turn, cause noise currents $i_2'$ in the output circuit. The resultant output noise current is $i_2''$. If a capacitive feedback admittance is present, this results in further noise currents $i_2'''$. By suitable choice of input and feedback admittances it is possible to make $i_2''$ and $i_2'''$ largely compensate each other.

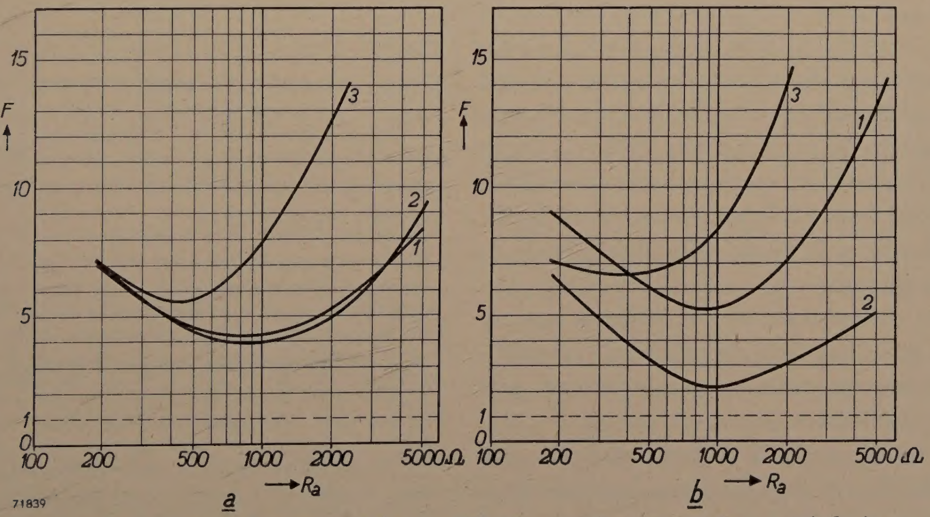

Fig. 11. Noise factor $F$ measured for an EF 50 pentode connected as a triode in a grounded-cathode circuit at a wavelength of 7 m, as a function of the reflected aerial resistance $R_{\mathrm{a}}$ in the input circuit. (*a*) applies to the case where no non-electronic feedback admittance is present in the circuit, (*b*) for the case where this is present, i.e. when the capacitance between grid and anode is not neutralised. Curves *1*: tuned input circuit, i.e. with real value of the input admittance $Y_1$; curves *2*: capacitively detuned input circuit ($Y_1$ practically wholly capacitive); curves *3*: inductively detuned input circuit ($Y_1$ practically wholly inductive). The noise factor always has a minimum at certain values of the reflected aerial resistance. The minimum is smallest for non-neutralised feedback and capacitive detuning (curve *2* of *b*).

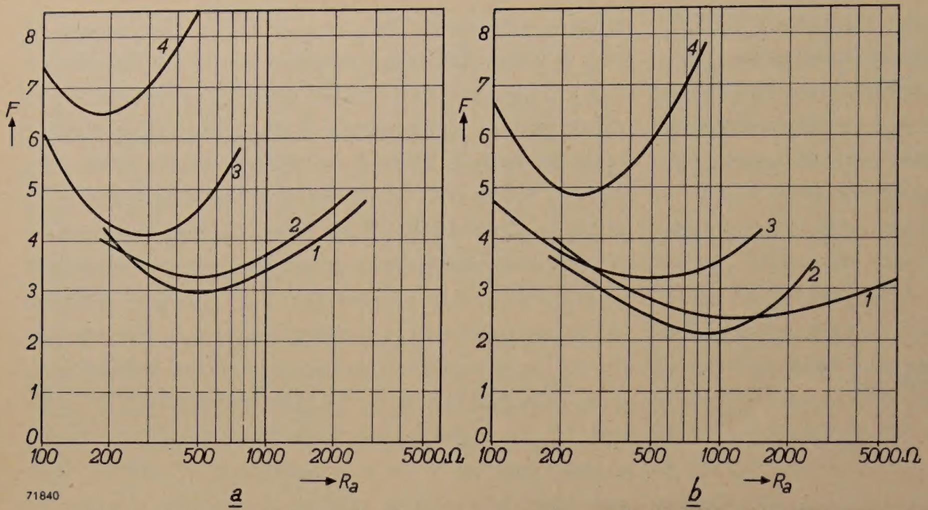

Fig. 12. Noise factor $F$, measured for an EF 42 pentode connected as a triode in grounded-cathode circuit at a wavelength of 7 m, as a function of the reflected aerial resistance $R_a$. (*a*) applies to a tuned input circuit (input impedance $Y_1$ real), (*b*) applies to an input circuit detuned by 5 pF ($Y_1$ practically wholly capacitive). Curves *1*: no feedback admittance (neutralised); curves *2*, *3* and *4*: capacitive feedback with $C_t$ = 0.5, 1, and 1.5 pF respectively. The minimum noise factor is lowest with capacitive detuning and 0.5 pF feedback admittance (curve *2* of *b*).

the reflected aerial resistance $R_a$ with tuned input circuit ($Y_1$ real). There is again, as with the grounded grid circuit, a minimum noise figure, in this case at $R_a \approx 1000\ \Omega$. Curve *2* gives the variation of the noise as a function of $R_a$ for a capacitively detuned input circuit ($Y_1$ capacitive apart from the small component $t^2/g$), and curve *3* for an inductively detuned input circuit ($Y_1$ practically wholly inductive). It is seen that, in agreement with the above, the smallest minimum noise factor is obtained with a capacitively detuned input circuit; inductive detuning increases the noise factor considerably (and also shifts the minimum).

The noise factor can be still further reduced, as will be finally shown, by making use of a feedback admittance $Y_{12}$ between anode and grid. If this is made purely capacitive, $\Delta V_g$ produces a further noise current $i_2'''$ in the output circuit, leading 90° with respect to $\Delta V_g$ [13]). The sum of all the noise vectors in the output circuit is thus further reduced. An example of this is given in fig. 11*b*, which concerns the same valve as in fig. 11*a*. Now, however, neutralisation is omitted, so that a capacitive feedback is, indeed, present between anode and grid. The curves *1*, *2* and *3* of fig. 11*a* now change to the corresponding curves *1*, *2*, and *3* in fig. 11*b*. It is clear that the presence of the feedback and the use of capacitive detuning does in fact markedly reduce the minimum noise factor compared with that without feedback.

According to fig. 10 it should therefore be possible, theoretically, to build a noise-free amplifier. A fuller examination shows, however, that with an optimum adjustment the signal gain drops below unity. Moreover, the correlation between the various noise currents is smaller in practice than it is in theory. It is clear, however, that a quite considerable improvement in noise factor is attainable.

In conclusion, *fig. 12* gives measurements of the noise factor with an EF 42 pentode used as a triode at a wavelength of 7 m. In this case not only is the tuning of the input circuit varied, but also the feedback capacitance; this was effected by shunting various capacitors $C_t$ across the neutralising inductance. At $C_t = 0$, there is no feedback admittance; values of $C_t$ differing from zero produce a corresponding feedback capacitance. Fig. 12*a* applies to an amplifier with tuned input circuit ($Y_1$ real), and fig. 12*b* to a capacitively detuned input circuit (by 5 pF) ($Y_1$ practically wholly capacitive). The capacitive detuning of the input circuit clearly has a favourable influence on the noise factor. A small capacitive feedback admittance gives still further improvement; a large capacitive feedback has the opposite effect, however, in the same way as an inductive feedback admittance (not shown in the figure). From these figures it is seen that a suitable choice of feedback admittance is of great importance for rendering the noise factor as small as possible. This influence of the feedback admittance in grounded-cathode circuits is often denied in current opinions.

**Summary.** This article, which is a continuation of an article with the same title published recently, deals with the noise of a triode at very high frequencies. It is shown that in the input circuit of a triode, at frequencies of the order of the reciprocal of the transit time of the electrons, a new type of noise is produced, the induced input noise. This is caused by the fact that noise currents which are induced in the external circuit leads give rise to grid voltage fluctuations. The induced input noise may influence the output circuit noise if the valve circuits have finite impedances. A triode can be used in several types of circuit, e.g. in grounded-grid and in grounded-cathode circuits. In both types of circuit a certain amount of compensation may occur in the various noise currents at the valve input, since phase relations exist between the noise currents in both inter-electrode spaces. The behaviour of triodes as amplifiers in both circuits mentioned is discussed. The sensitivity of the amplifier depends on the noise in the output circuit, and is expressed by a specially defined noise factor.

The influence which a feedback admittance between input and output of the valve may exert is illustrated.

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[13]) See, for example, the references cited in notes [3]) and [6]).